THE PRACTICAL MAGAZINE WITH THE PROFESSIONAL APPROACH

July/August 1988

Elektor Electronics

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Chrominance-locked clock generator
Microprocessor-controlled radio synthesizer
GaAs FET converter for 23 cm ATV
Transmission & reception of RTTY

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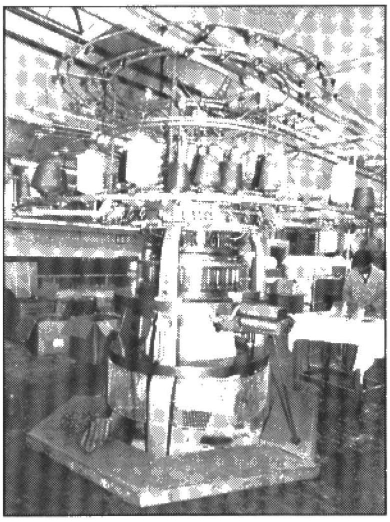

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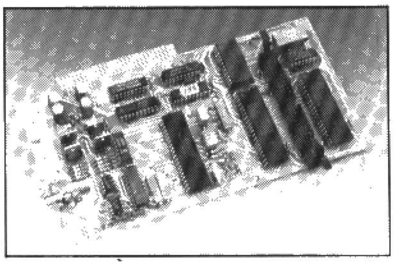

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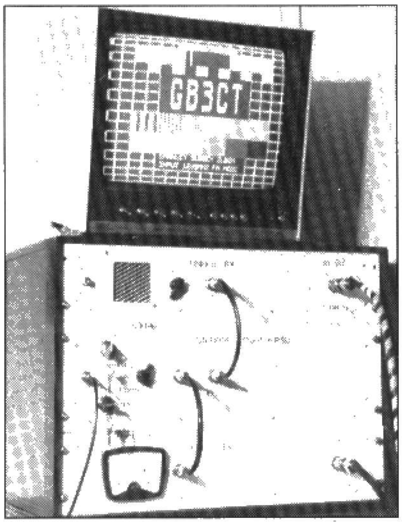

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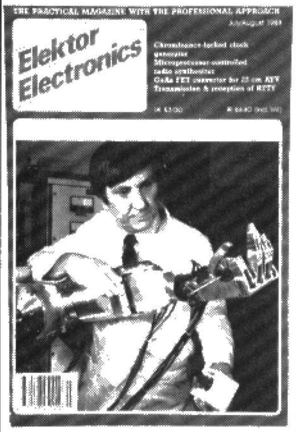

Front cover

A robotic arm is put through its paces before taking its place in one of Britain's industrial research establishments. The hand, with almost as many joints as a human hand, and claimed to be just as versatile, is capable of gripping up to 5 kg (11 lb). Southampton University has been researching robotic hands for more than 15 years, producing a number for medical uses. Some of these are not only natural in appearance, but also have a sense of 'touch'.
This particular arm, which has separate controls for the shoulder, forearm, elbow and hand, has been designed for use with dangerous materials in special industrial environments. All the movements are made by a series of powerful miniature electric motors. Ten microcomputers are hooked up between the arm and joystick, effectively forming the 'brain'. It is said to be capable of moving in a 180 degree arc in one second and to have a repeatable accuracy of ±0.1 mm.

In our September issue:

- Microcontroller-driven power supply — Final part
- 64 K RAM for MSX
- Shielding computers with metal-coated glass
- Microphone/line active filter/driver
- Inductance meter
- Microprocessor-controlled radio synthesizer — Final part

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ELECTRICAL SAFETY

In Britain, as in most western countries, we have strict legal requirements to ensure that mains-operated equipment is safe as regards electric shock and fire caused by undue rises in temperature. The relevant laws and regulations apply equally to proprietary and home constructed equipment and installations.

Also, in Britain we have one of the safest domestic electricity supplies in the world. Yet, an average of 40 people die each year in this country from electricity-related accidents, while a further 2,000 need hospital treatment. Although these figures compare well with those of most other countries, they are not good.

It is probably because of the high degree of inherent safety and the relevant legal requirements that most of us take safety for granted and are no longer (at least consciously) aware of the risks of injury or death from electric shock or fire caused by electrical faults. Invariably, such faults are man-made.

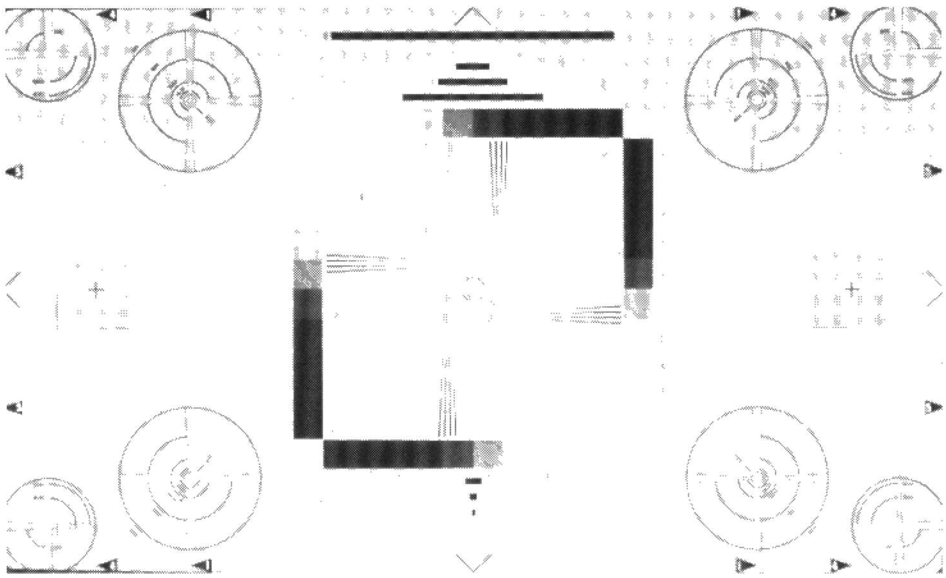
To increase people's awareness, the Department of Trade and Industry and the Electrical Installation Equipment Manufacturers' Association (EIEMA) have recently launched an Electrical Safety Awareness Campaign. One of the aims of this campaign is to persuade people to use Residual Current Devices (RCDs) to reduce the risk of injury or death.

Basically, an RCD is capable of detecting an electric current that flows anywhere other than the circuit intended for it, and switches it off rapidly so that it can prevent electrocution or an undue rise in temperature.

The Wiring Regulations of the Institution of Electrical Engineers already require the use of RCDs within an installation where hazards are recognizably greater than normal, such as electrical equipment used outdoors.

Since an RCD provides protection that supplements the basic measures and offers a degree of safety against the inevitable wear and tear or any misuse, abuse or degradation of the inbuilt safety of electrical installations and equipment, it is to be hoped that people will start making far greater use of it, particularly, but not only, with electrical musical instruments. An RCD Information Bureau has been established that may be contacted by telephoning the operator and asking for 'Freephone RCD Devices'.

CHROMINANCE-LOCKED CLOCK GENERATOR



A remarkably simple solution is offered to a problem almost any constructor of a test card or callsign generator, logomat, graphics card, or any other video equipment must have been faced with at some time: phase synchronicity between clock pulses in the system and the chrominance subcarrier.
A PLL circuit is described that enables deriving the TV line and field frequency, and a number of other useful signals, from the chrominance frequency, 4.433 MHz. A mystery unravelled!

by J. C. Stekelenburg PE1FYZ

In video equipment, the beneficial effects of phase-locking the central clock oscillator to the chrominance subcarrier are mainly the elimination of annoying digital interference and colour cross-patterning. The improvement can be noticed in the well-known colour bar test chart, in which colour transitions become sharply defined rather than blurred with bands of spurious lines and randomly moving coloured spots while longer lines are moving slowly and diagonally or horizontally across the screen.
In professional video systems and studios, complex equipment is available to ensure that all TV synchronization signals have a fixed phase relationship with the chrominance frequency, as set forth in the relevant CCIR specifications.

This article demonstrates how some thinking on the technical background of the PAL (*Phase Alternation Line*) and NTSC (*National Television System Committee*) TV systems, and a comparison between these in respect of possible interference, leads up to simple computer-assisted arithmetic and, finally, the design of a circuit that achieves the above objective of providing chrominance-locked, standard clock frequencies for digital video generators.

Choice of the chrominance subcarrier frequency

The PAL TV system is based on double-sideband modulation of the picture colour information onto a subcarrier of 4.433 MHz. This system is basically similar to that used for NTSC TV. The Y (luminance) signal is obtained by adding the three primary colours, red (R), green (G) and blue (B), in proportion, as

$$Y=0.3R+0.59G+0.11B$$

The degree of luminance of each individual pixel determines its brilliance, black corresponding to minimum, and white to maximum luminance. For a monochrome TV set, the luminance signal is sufficient for producing a picture. A colour receiver, however, needs the three primary colours for mixing to give each pixel on the screen the correct colour. The colour receiver finds two modulated signals, R−Y and B−Y adjacent to the 4.433 MHz subcarrier. From these, the R, G and B signals are obtained by means of a number of

simple operations involving subtraction and addition. The R–Y and B–Y signals are quadrature-modulated on the 4.433 MHz carrier, so that the instantaneous phase provides a measure for the colour of a pixel, and the amplitude for the colour saturation. Since the colour subcarrier is found within the luminance band (0. . .5 MHz), its sidebands will become visible as a pattern of thin lines. In the NTSC TV system, this undesirable effect is minimized by using a colour subcarrier frequency that is an odd multiple of the line frequency. This gives rise to a dot pattern which is far less conspicuous and annoying than the line pattern that would be formed in the PAL picture (see Fig. 1).
In the PAL system, the phase of the modulated R–Y signal is inverted for every line in the picture. This causes two sidebands adjacent to the 4.433 MHz subcarrier, at an offset corresponding to half the line frequency. When the subcarrier frequency is chosen such that it is an odd-numbered multiple of *half* the line frequency, B–Y information will result in a dot pattern, and R–Y information in a line pattern. To avoid this, the chrominance subcarrier frequency is an odd multiple of the line frequency divided by four (*quarter-offset*). Time-averaging of the remaining interference on a raster-by-raster basis is further achieved by adding 25 Hz (raster frequency) to the colour subcarrier, so that cross-interference between luminance and chrominance is least noticed. Summarizing the above, the optimum frequency of the chrominance subcarier, f_{chr}, becomes:

$$f_{chr} = (15{,}625/4)\cdot 1135 + 25$$
$$= 4{,}433{,}618.75 \text{ Hz}$$

corresponding to the line frequency multiplied by 283.7516. The deriving of the line frequency from the chrominance frequency would, therefore, require dividing this by 283.7516. This is impossible by electronic means, which only allows dividing by integers. Subtraction of 25 Hz, division by 1135 and subsequent multiplication by 4 is also very complex in terms of electronics. Reasonable accuracy is, however, obtained by approximation of the denominator, 283.7516.

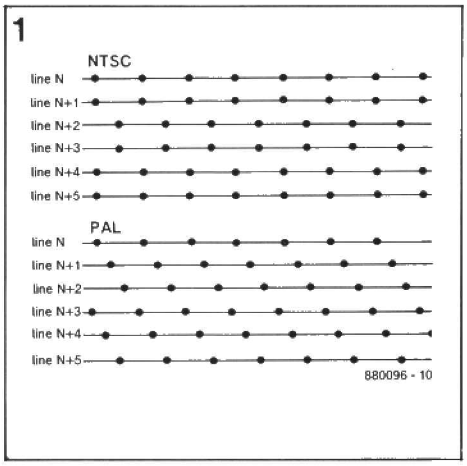


Fig. 1. Comparison between PAL and NTSC: luminance patterns caused by chrominance information.

Table 1.

```
1 REM chrominance-locked clock generator
5 REM successive approximation of clock divisors
10 K=4433618.75
20 INPUT''Enter output frequency '',F
30 H=K/F
40 P=H/(H-1)
50 B=1
60 FOR X=1 TO 1000
70 Y=P*X
80 A=ABS(Y-RND(Y,-1)/X
90 IF A>=B; NEXT X
100 IF A<B; B=A : D=RND(X*P,-1) : E=D-X : PRINT X,D,E
110 R=K/D*E : U=K/D
120 S=R/256-15625
130 PRINT R,S,U
140 NEXT X
```

Table 2.

Objective: f_0 = 4 MHz

x	divisor *d*	divisor *e*	real output frequency [Hz]	deviation from 15,625 [Hz]	phase comparator frequency [Hz]
1	10	9	3,999,256.875	−38.05908	443,361.875
3	31	28	4,004,558.871	17.80809	143,019.9597
4	41	37	4,001,070.579	4.1819501	108,137.0427
5	51	46	3,998,950.245	−4.1006052	86,933.70098
9	92	83	3,999,895.177	−0.4094664	48,191.50815
31	317	286	4,000,047.2	0.1843761	13,986.17902
40	409	369	4,000,013.004	0.0507978	10,840.14364
49	501	452	3,999,991.367	−0.0337217	8,849.538423
89	910	821	4,000,001.092	0,0042657	4,872.108516
227	2321	2094	3,999,998.993	−0.0039341	1,910.219194
316	3231	2915	3,999,999.584	−0.0016247	1,372.21255
405	4141	3736	3,999,999.916	−0.0003293	1,070.663789
899	9192	8293	4,000,000.032	0.000125	482.3345028

Wanted: two denominators

Many digital video circuits have a central clock oscillator that runs at a multiple of 1 MHz. This is so arranged because the line frequency is then readily obtained with the aid of binary counters/dividers. For instance, when a clock of 4 MHz is available, 15,625 Hz is obtained by division by 256 ($=2^8$). The question is now: how can we relate 4,433,618.75 Hz to 4,000,000 Hz? The answer can be provided by a computer, programmed to find two integer denominators: one, *d*, for the chrominance frequency, and another, *e*, for the clock frequency. In other words, if the chrominance frequency is divided by *d*, and the result of this division is multiplied by *e*, 4,000,000 MHz should be obtained.
The BASIC computer program listed in Table 1 gave the results summarized in Table 2. Denominators 910 for *d* and 821 for *e* were found to yield a reasonable approximation of the target frequency whilst giving a practicable operating frequency for the phase comparator in the PLL to be designed. Also, these denominators allow relatively simple divider circuits to be used. The final deviation from 4,000,000 MHz is virtually negligible at +1.092 Hz.

Practical circuit

The above considerations lead up to the block diagram of Fig. 2. It is seen that 4 MHz is divided by multiples of 2 to obtain commonly used frequencies in digital video circuits. The chrominance frequency is multiplied by two to give 8.86 MHz, which is frequently required as a clock signal for IC-based colour generators.

Figure 3 shows the circuit diagram of the chrominance-locked clock generator, which is essentially a discrete phase-locked loop designed around commonly available parts.

The 4.433 MHz crystal oscillator is set up around gate N_5, whose output signal is fed to buffer N_1 and counter IC_3. This, together with bistable FF_1, divides the chrominance frequency by 910 to give 4.8721085 kHz. IC_3 counts 909 periods of the clock signal, while FF_1 delays 1 period (approx. 226 ns) during the resetting of the counter, giving the required divisor and allowing sufficient time for IC_3 to reset all internal bistables.

The 4 MHz L-C oscillator is a varicap-controlled Colpitts type set up around T_1, with T_2 and Schmitt-trigger gate N_4 acting as a buffer to obtain a digital compatible output signal.

Division by 810 in IC_4 is achieved in a manner similar to that in IC_3-FF_1 as discussed.

Gate network N_2-N_6-N_7-N_3 forms the phase comparator, while R_4-C_3 forms the loop filter with a roll-off of 7.2 Hz. The natural frequency of the PLL is 12.96 Hz, while the hold range is of the order of 150 kHz.

Construction and setting up

The circuit is fairly uncritical in respect of construction, and is simple to build on a small piece of Veroboard. Connections in the 4 MHz and 4.433 MHz oscillators should be as short as possible though, and due attention should be paid to decoupling of the positive supply line.

The only component that requires further discussion here is L_1. In the prototype, good results were obtained with a Neosid Type 7A0 inductor assembly. The required inductance of about 22 μH was achieved by winding 60 turns of 0.2 mm dia. enamelled copper wire onto the former. Do not use ready-made 4.433 MHz inductors with a built-in parallel capacitor, since this is often too large to ensure the relatively high L-C ratio required in this application. The 4.433 MHz crystal was a type salvaged from a colour TV chassis.

After building the circuit, it is recommended to commence testing the 4 MHz oscillator by temporarily breaking the PLL control loop. Disconnect R_4 from pin 4 of N_3. This enables checking the operation of the 4 MHz oscillator with the aid of an external tuning voltage obtained from the wiper of a po-

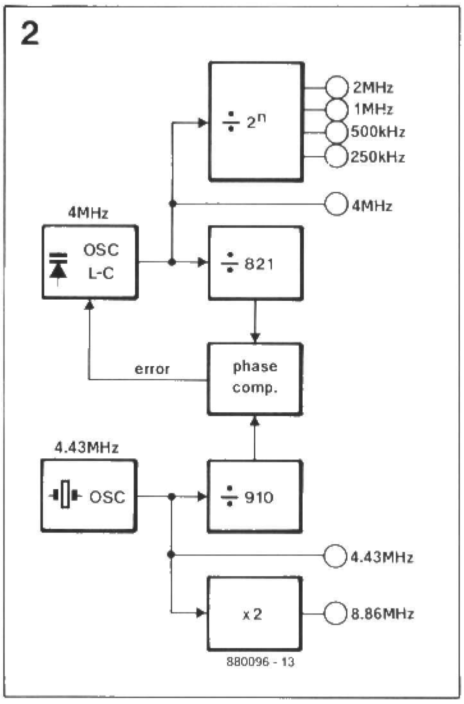


Fig. 2. Block diagram of the chrominance-locked clock generator for video systems.

Parts list

Resistors (±5%):
R_1 = 10M
R_2 = 120K
R_3;R_7;R_{11};R_{12} = 10K
R_4;R_6 = 100K
R_5 = 560K
R_8 = 12K
R_9 = 1K0

Capacitors:
C_1 = 100p
C_2 = 40p trimmer
C_3 = 220n
C_4;C_5;C_9;C_{10};C_{13} = 47p
C_6 = 68p
C_7;C_8;C_{11} = 100p
C_{12} = 1n0
C_{14};C_{15} = 100n

Inductors:
L_1 = winding details are given in the text. Neosid Type 7A0 (Neosid Limited • Icknield Way West • LETCHWORTH SG6 4AS. Telephone: (0462) 481000. Telex: 826405. Contact Mr. E. Adcott). Neosid inductors are also available from C-I Electronics.
L_2 = 13μH with centre tap (see text).
L_3 = 7μH with centre tap (see text).

Semiconductors:
D_1 . . . D_{11} incl. = 1N4148
D_{12} = BB105
D_{13};D_{14} = AA119
IC_1 = 4093
IC_2 = 4001
IC_3;IC_4 = 4040
IC_5 = 4013
IC_6 = 4516
T_1;T_2 = BF199

Miscellaneous:
X_1 = quartz crystal 4.433 MHz.

It is regretted that a printed-circuit board for this project is not available.

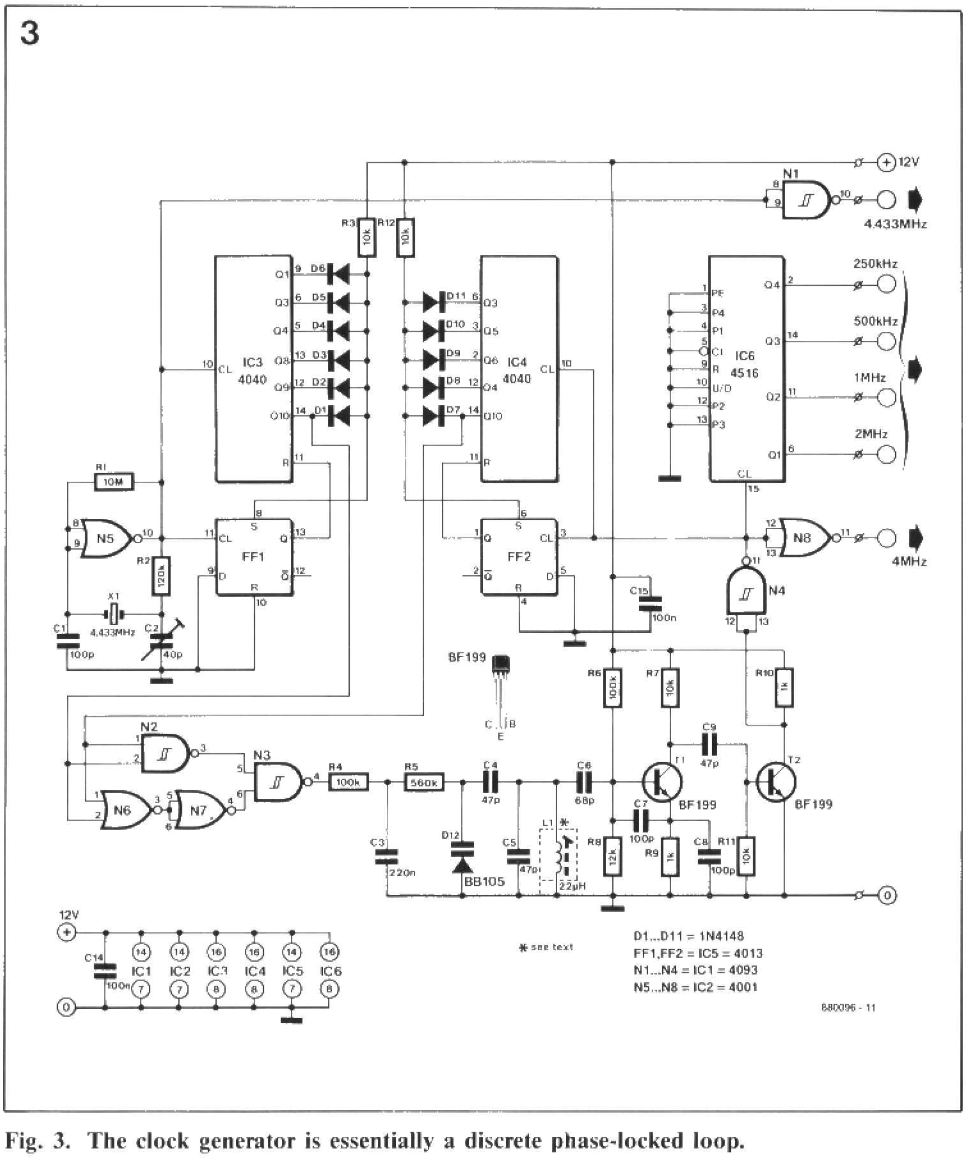


Fig. 3. The clock generator is essentially a discrete phase-locked loop.

tentiometer connected between +12 V and ground. To begin with, adjust L_1 so that oscillation is achieved around 4 MHz. Check that the oscillator can be tuned with the potentiometer. The output of N_4 should supply CMOS-compatible clock pulses. It is essential that these pulses have the full CMOS swing of about 12 V_{pp} when the oscillator is tuned around 4 MHz. Do not add too much capacitance to the parallel tuned circuit when its resonance frequency is found to be too high: instead, ensure more inductance by increasing the number of turns on L_1.

Next, adjust trimmer capacitor C_2 to give 4.43362 MHz at the buffered output. Measure the frequency at pin 2 of N_2 — this should be 4.8721085 kHz. Similarly, measure the frequency at pin 1 of N_2 to check the operation of IC_4-FF_2. Tune the oscillator to obtain about 4.8 kHz here.

When these tests check out, it is time to close the loop by removing the potentiometer, and connecting R_4 to the output of N_3. Connect the frequency meter to the 4 MHz output. Some re-adjustment of L_1 may be required to get the PLL to lock.

The oscilloscope photographs of Fig. 4 may be used as guidance if difficulties are encountered in the setting up. The upper two traces show the 4.8 kHz signals at the inputs of the phase comparator, i.e., pins 1 and 2 of N_2 (or N_6), the lower trace the phase comparator output (pin 4 of N_3). Although the latter signal is different in the photographs, the PLL was locked in both conditions, with only L_1 set differently within the hold range of the oscillator. It is clearly seen that the phase comparator is essentially an exclusive NOR function: the output goes low only when the two input signals are different.

The operation of the PLL can be checked by carefully adjusting L_1 while monitoring the phase comparator output with an oscilloscope. It will be found that the PLL loses lock when the pulses become significantly narrower than those in the lower trace of Fig. 4b. When the PLL is locked, L_1 can be adjusted over a small span while the output frequency remains stable at 4.000002 MHz (7-digit resolution).

Finally, switch the power to the circuit on and off a few times to verify that the PLL starts and locks properly. All drift on the 4 MHz output is, of course, caused by drift of the quartz crystal frequency. It is, therefore, recommended to make the final adjustment of C_2 and L_1 after a warming-up period of about 10 minutes.

4

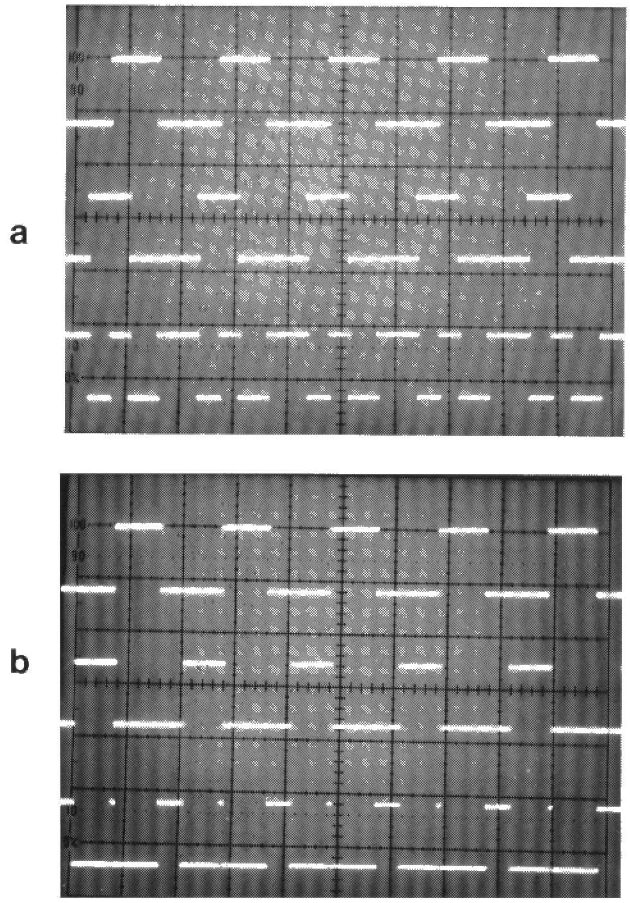


Fig. 4. Oscillograms showing the operation of the XNOR phase comparator.

5

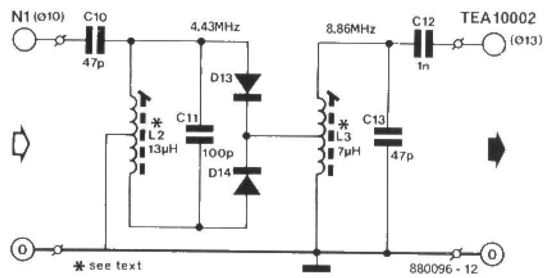


Fig. 5. Frequency multiplier circuit for driving the Type TEA1002 colour generator.

Multiplier for TEA1002

The circuit described is used by the author as part of a digital test chart and call-sign generator for amateur television. The system incorporates a TEA1002 colour generator chip (Ref. [1]) which requires an input signal of 8.86 MHz. The circuit of Fig. 5 multiplies the buffered crystal oscillator output of the chrominance oscillator by two to obtain this frequency. The multiplier is essentially a double-phase rectifier with a parallel-resonant L-C output filter. Suggested diode types are AA119 or OA95 (in any case, germanium types should be used). L_2 and L_3 are wound as 30 and 20 turns respectively of 0.2 mm dia. enamelled copper wire, with a centre tap. It is also possible to use ready-made inductors provided they are known to have a centre tap and the correct inductance (use a grid-dipper to check the in-circuit resonance frequency). Both L_2 and L_3 are simply peaked for maximum amplitude of the 8.86 MHz output signal. *Bu*

Reference:

[1] Video combiner. *Elektor Electronics* February 1984, p. 2-36 ff.

For further reading:

Principles of PAL Colour Television. H. V. Sims, London Iliffe Books Ltd. 1969.
Phase-lock techniques. F. M. Gardner, John Wiley & Sons Inc. 1966.

MICROPROCESSOR-CONTROLLED RADIO SYNTHESIZER — 1

**The addition of a microprocessor-controlled synthesizer to a continuously-tuned receiver greatly improves tuning accuracy and provides several additional facilities that have become available in recent years.
The versatile synthesizer described has a 6-digit LCD or LED display and a 16-position keyboard which allows direct frequency entry, channel or frequency increment or decrement, as well as the storing and recalling of 30 frequencies. The MW, SW and FM band are covered each with a choice of IF offsets.**

by P. Topping

Most recently designed quality radios employ synthesized local oscillators controlled by a microprocessor. These complex designs should not discourage the advanced home constructor, however, as components are currently available which enable similar facilities to be either incorporated in individual designs, or added to existing radios.
Synthesis of the local oscillator (LO) in a superheterodyne receiver provides many advantages over the more traditional mechanical tuning. The main benefits are improved tuning accuracy, stability and the possibility to store often used frequencies. Accuracy and stability result from the fact that the local oscillator is phase-locked to a reference crystal oscillator. Before synthesizers became available, crystals were used to obtain a good degree of accuracy. This has the disadvantage of requiring a separate crystal for each frequency. Using a phase-locked loop (PLL) synthesizer, similar performance can be achieved at an unlimited number of frequencies from only one crystal. Accurate, drift-free, tuning is particularly important for stand-by use of a receiver when nobody is on hand to provide fine tuning.
A synthesizer can be incorporated into almost any receiver simply by replacing the tuning capacitor with a variable capacitance diode (varicap) as shown in Fig. 1. The voltage biasing this varicap is supplied by the synthesizer, which thus takes over the RF tuning. A simpler solution is to retain the RF function of the existing tuning control as a preselector to avoid tracking problems in multiband designs. The current trend is to eliminate front-end tuning altogether, and employ only a wideband filter between the RF input and the first mixer.

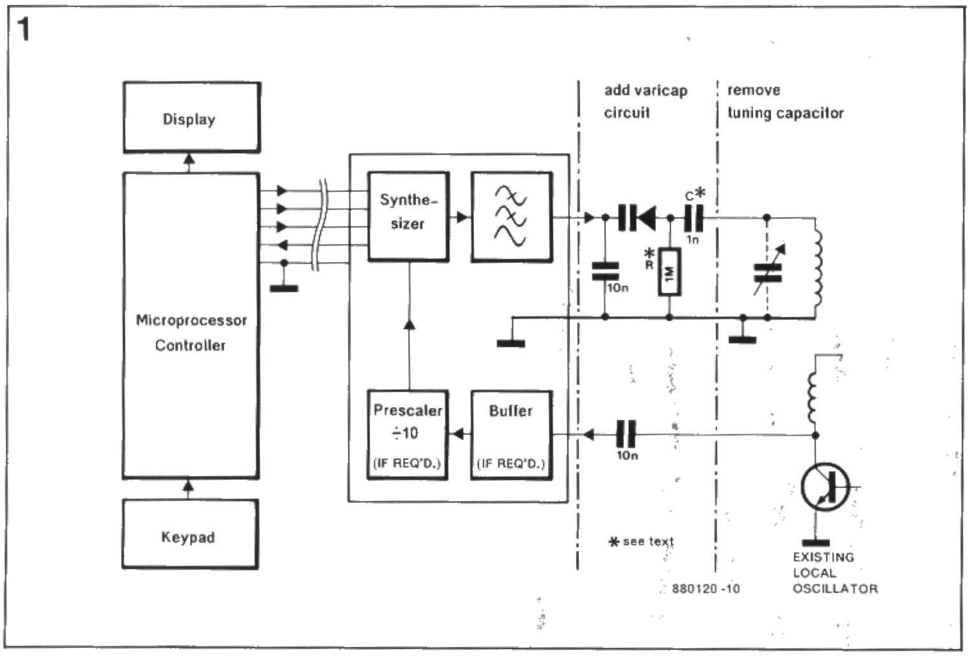


Fig. 1. A local-oscillator synthesizer can easily replace mechanical tuning to provide crystal-controlled accuracy and many other improvements in performance and convenience.

Synthesizer MC145157

The Type MC145157 CMOS synthesizer from Motorola is one of a series offering a variety of options including serial or parallel interfacing, and single or dual modulus prescaling. In the synthesizer described here, only single modulus

MULTIBAND RF SYNTHESIZER

Features:

- Coverage of MW, SW and VHF FM bands.
- Variable step size in accordance with station spacing.
- CMOS design allows battery back-up while minimizing power consumption and RF interference.
- 11 switch-selected bands with a variety of IF offsets, including nought.
- Power-down mode for display and processor.
- Easy to operate keyboard.
- RIT (receiver incremental tuning) control provided.
- Choice of three 6-digit displays: 7-segment LED, static LCD or multiplexed LCD.
- Memory function for up to 30 often-used frequencies.
- Last used frequency automatically recalled at receiver power-on.
- Simple to incorporate in almost any general-purpose receiver.
- Direct synthesis up to 16 MHz without prescalers.
- Prescalers for up to 60 MHz and up to 150 MHz.
- IF offset can be customized to indivual requirement.

2

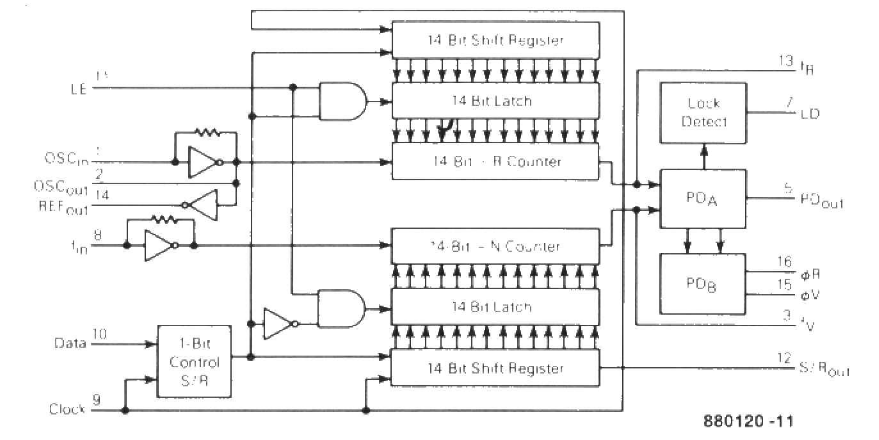


Fig. 2. The MC145157 synthesizer includes two 14-bit shift registers, one each for the reference divider (top) and the variable (LO) divider. Their loading is controlled by the value of a trailing 15th bit. The outputs of these dividers are compared in respect of phase to control the frequency of the local oscillator.

3

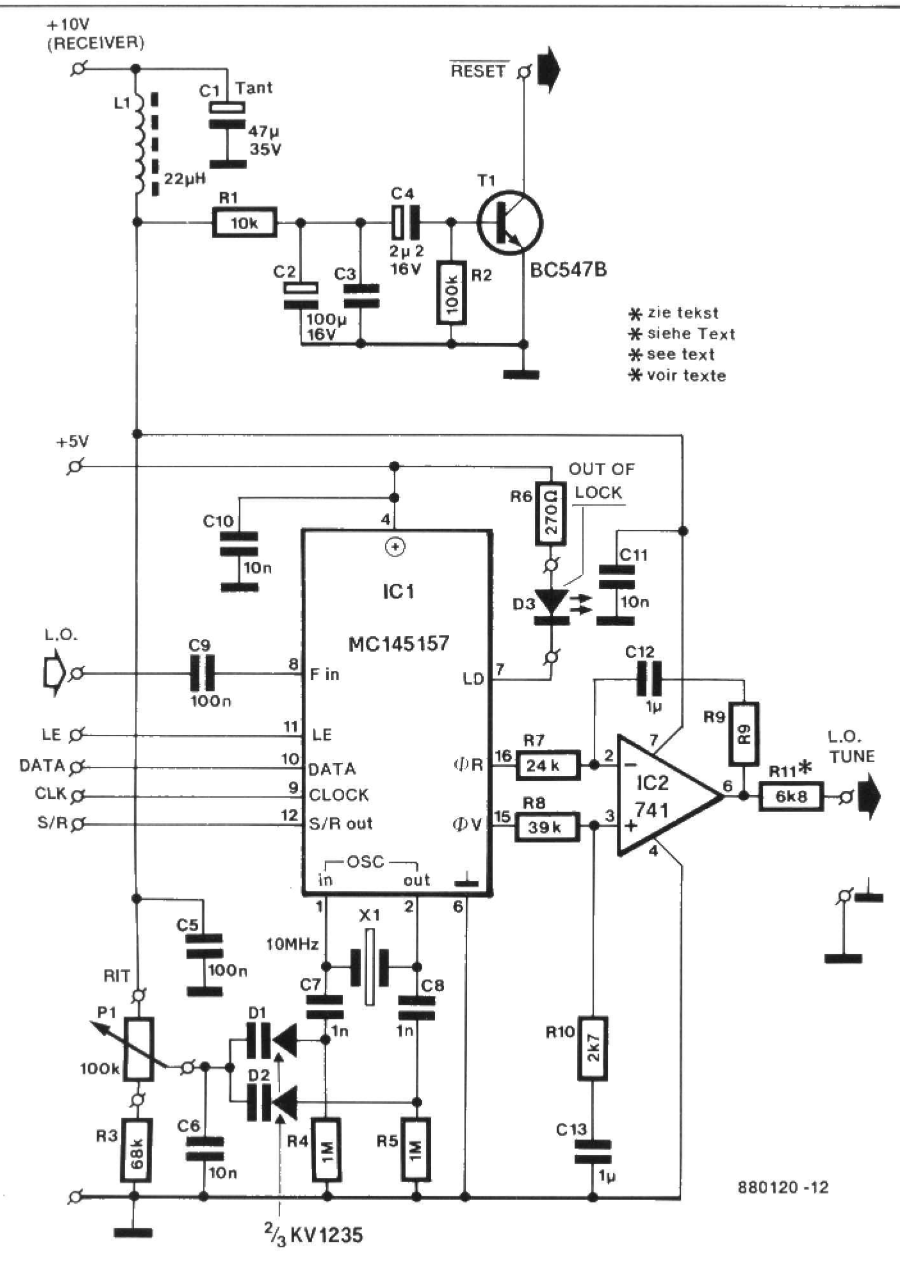


Fig. 3. The synthesizer module contains the MC145157 phase-locked loop chip and its opamp-based active low-pass filter. It also includes a RIT control and a microprocessor reset circuit, which makes it possible to resend the last used frequency in stand-by applications.

prescaling is used. Serial interfacing was chosen to minimize the number of interconnections between the synthesizer and the microprocessor.

The block diagram of the MC145157 is shown in Fig. 2. There are two 14-bit long counters, which are loaded by shift registers, starting with the most significant bit (MSB). After loading the 14 databits, a 15th control bit is loaded, and the information is transferred to the selected latch using LE (latch enable). If the control bit is a logic one, the reference divider latch is loaded; if it is a logic zero, the variable (LO) divider latch is loaded.

The reference counter divides the crystal oscillator down to the reference frequency (in this case 1 kHz), at which the comparison is made with the (also divided down) local oscillator frequency. The error signal from the phase comparator is filtered, and forms the tuning voltage for the local oscillator. The numbers chosen as the divide ratios determine the frequency at which this oscillator stabilizes. The equation below shows the relationship between the various frequencies, where P is the LO prescaler, N is the reference divider ratio, and Q is the LO divider ratio. The received frequency can be changed by altering the LO divide ratio. The microprocessor takes care of the decimal to binary conversion, IF offset, and the other required arithmetic.

$$f_{LO} = RF + IF = P(f_{Xtal}/N)Q$$

The practical application of the MC145157 is shown in the circuit diagram of Fig. 3. The output signal of the synthesizer's 10 MHz crystal oscillator is divided down by 10,000 to obtain the reference frequency at which the phase comparator operates. A compromise is required when deciding on this frequency. Filter design is relaxed by choosing a high reference frequency, but the disadvantage is that the minimum step size of the synthesizer is determined by the reference frequency, as the smallest change which can be made results from a change of 1 on the LO divide ratio (see the above equation). A reference of 1 kHz is a reasonable compromise for most broadcast receivers.

The MC145157 is specified to operate up to 20 MHz, so prescaling is required on FM (VHF) and SW. Shortwave band divide-by-5 prescaling is used, and for FM divide-by-10. This increases the minimum step size to 10 kHz on FM, which is ideal for this band, and to 5 kHz on SW, which is suitable for most broadcast receivers but too large for some shortwave applications. Fortunately, however, this can be alleviated by the use of an RIT (*receiver incremental tuning*) control, formed by external potentiometer P_1. The low-IF (455, 468 and 470 kHz) SW bands do not use prescal-

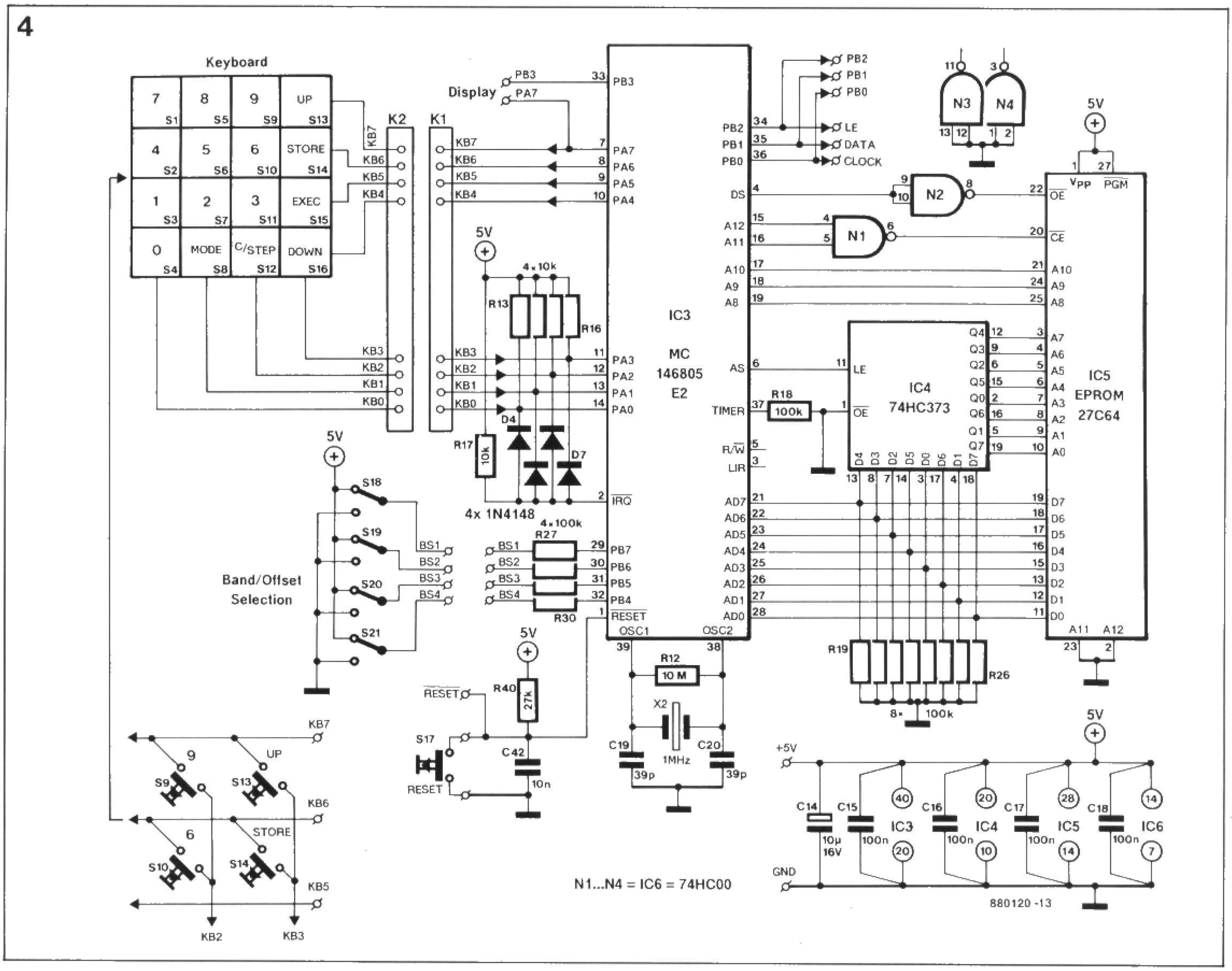


Fig. 4. Circuit diagram of the microprocessor-based controller and the keyboard.

ing, and thus have a step of 1 kHz, but a maximum frequency of just under 16 MHz. (2^{14} – IF).

Complex multiple loops are used in some commercial designs to achieve better resolution, but this can also be achieved with the previously mentioned RIT control. With reference to Fig. 3, the adjustment is made by slightly changing the synthesizer's reference frequency. This can be accomplished by replacing the usual combination of a fixed and a trimming capacitor on the crystal pins of the MC145157 with varicap diodes (D_1-D_2). Adjustment is thus by a direct voltage, taken from the wiper of potentiometer P_1, which can be fitted remote from the synthesizer.

This type of adjustment necessarily gives a control range which depends on the tuned frequency, but the relatively high IF ensures that this is not too significant. For instance, using an IF of 10.7 MHz, the circuit shown gives the required range of ±2.5 kHz at the lower end of the SW band (1.6 MHz), and just over twice this range at about 15 MHz. If an RIT control is not required, pins 1 and 2 should have a 47 pF capacitor and a 30 pF trimmer to ground respectively, the trimmer being adjusted to provide a reference of 1 kHz. If a frequency meter is not available, this adjustment can easily be made by tuning into a strong broadcast of known frequency, and adjusting for optimum reception and symmetric off-channel response.

An important part of any phase-locked loop is the loop (low-pass) filter. The active filter set up around opamp IC_2 is driven from the double-ended phase detector output of the MC145157. An active filter has the advantage of increasing the available voltage swing beyond the supply rail (5 V) of the synthesizer chip. The supply voltage on the active filter determines the maximum voltage available to the varicap diodes in the RIT circuit — 10 V is suitable for the Type KV1235 triple varicap from Toko. The 10 V supply is conveniently taken from the receiver which incorporates the synthesizer.

The combination of active filter and double-ended phase detector outputs makes it simple to select the correct relationship between voltage and LO frequency. Usually, one side of the varicap diode is grounded, so that increasing the reverse voltage on it increases the frequency of the local oscillator. In some oscillator designs, however, the fixed side may be taken to the supply rail, so that increasing the tuning voltage decreases the frequency. With the filter design shown, the choice can be made simply by interchanging the connections to pins 15 and 16 on the MC145157.

Resistors R_7. . . R_{10} incl. may need to be adjusted empirically to stabilize the loop and eliminate any trace of the reference frequency from the output of the radio (remember that LO phase noise is demodulated together with the wanted signal).

Microprocessor and keyboard

The next module in the synthesizer is the microprocessor circuit — see Fig. 4. The microprocessor used is the CMOS Type MC146805E2 from Motorola, which offers powerful bit manipulation instructions, useful for this type of application. It also has a stand-by (*power-down*) mode in which the clock is stopped. This

has the double advantage of saving power in battery applications, and eliminating interference problems with the radio. When a key is pressed, the microprocessor 'wakes', performs the required function, and then goes back into the stand-by mode.
The MC146805E2 has a multiplexed bus for data and low-order addresses. This arrangement saves pins but requires an external address latch, IC_4, to interface with the system EPROM, IC_5. This is a CMOS Type 27C64 so that the whole system consumes only a few milliwatts when active, and a few microwatts in stand-by. Although the control program in the EPROM could have been accommodated in a 27C16 (2K×8) with room to spare, an 8 Kbyte EPROM was chosen because this is currently less expensive and generally better obtainable; the 2716, and its CMOS version, 27C16, is now rapidly becoming obsolete.
After performing the initialization routine at power-on, or following a reset, the microprocessor is programmed to switch lines PA4...PA7 of port A to output, and PAØ..PA3 to input. The output lines are set logic low before the CPU is software-switched to the power-down state. Any subsequent action on one of the keys $S_1 \ldots S_{16}$ incl. drives the processor's interrupt request ($\overline{IRQ}$) line logic low, ending the power-down state. Instructions in the EPROM cause the CPU to start scanning the keyboard with the aid of outputs PA4...PA7 and inputs PAØ...PA3, to determine which key was pressed, execute the appropriate command or load the pressed number on the keyboard, write serial data to the synthesizer via PBØ, PB1 and PB2, and update the display read-out via PB3-PA7, PBØ-PB2, or PB1-PB3-PA7 (the port lines used depend on the display type — this will be discussed later).
The EPROM-resident control program is located in address range 1800_H to $1FFF_H$. This is the top of the CPU's address space (it can address 8 Kbyte of memory), and includes the reset and interrupt vectors. These vectors reside between $1FF6_H$ and $1FFF_H$ with the program itself starting at $18ØØ_H$. The MC146805E2 microprocessor has a 112-byte on-chip RAM area in page zero. The CPU bus is demultiplexed by octal latch IC_4 using the address strobe (AS) pulse.
The 5 V supply to the controller should not be switched off if the station memories are to survive. The supply does not need to be regulated, and four 1.5 V zinc-carbon or Ni-Cd batteries will do. With the static LCD on, the current drawn in stand-by is about 50 μA, without it, less than 1 μA. Eight 100 kΩ pull-down resistors on the multiplexed bus lines of the CPU are used for ensuring minimum stand-by power dissipation. When the display is switched on, it will show random data, but will be written to when any key is pressed (use of the EXECUTE key restores the display to its previous data), or automatically by the reset circuit shown in Fig. 3.

Using the keyboard

The 16-key keypad performs the following functions:

0-9 These keys are used for both direct frequency entry and recalling (or storing) the ten frequencies available in each band.
UP Increment by one channel (5 KHz SW; 9 kHz MW; 50 kHz FM) or 1 kHz (10 kHz on FM, not applicable to 10.7 MHz SW).
DOWN Decrement by one channel (5 KHz SW, 9 kHz MW, 50 kHz FM) or 1 kHz (10 kHz on FM, not applicable to 10.7 MHz SW).
STORE Next key (0-9) stores current frequency at that key (indicated by a decimal point on the left-most digit).
CLEAR Clear display (direct frequency entry). Also toggles between channel steps and 1 kHz steps (indicated by a decimal point on the second digit from the right).
MODE Change between frequency and station mode.
EXECUTE Go to frequency, but stay in current mode.

The leftmost digit in the display indicates which mode is current:
- Display blank: *direct frequency entry mode*;
- Number: last station stored or recalled — *station mode*;
- Small square (LC display) or a single lit horizontal centre segment (LED

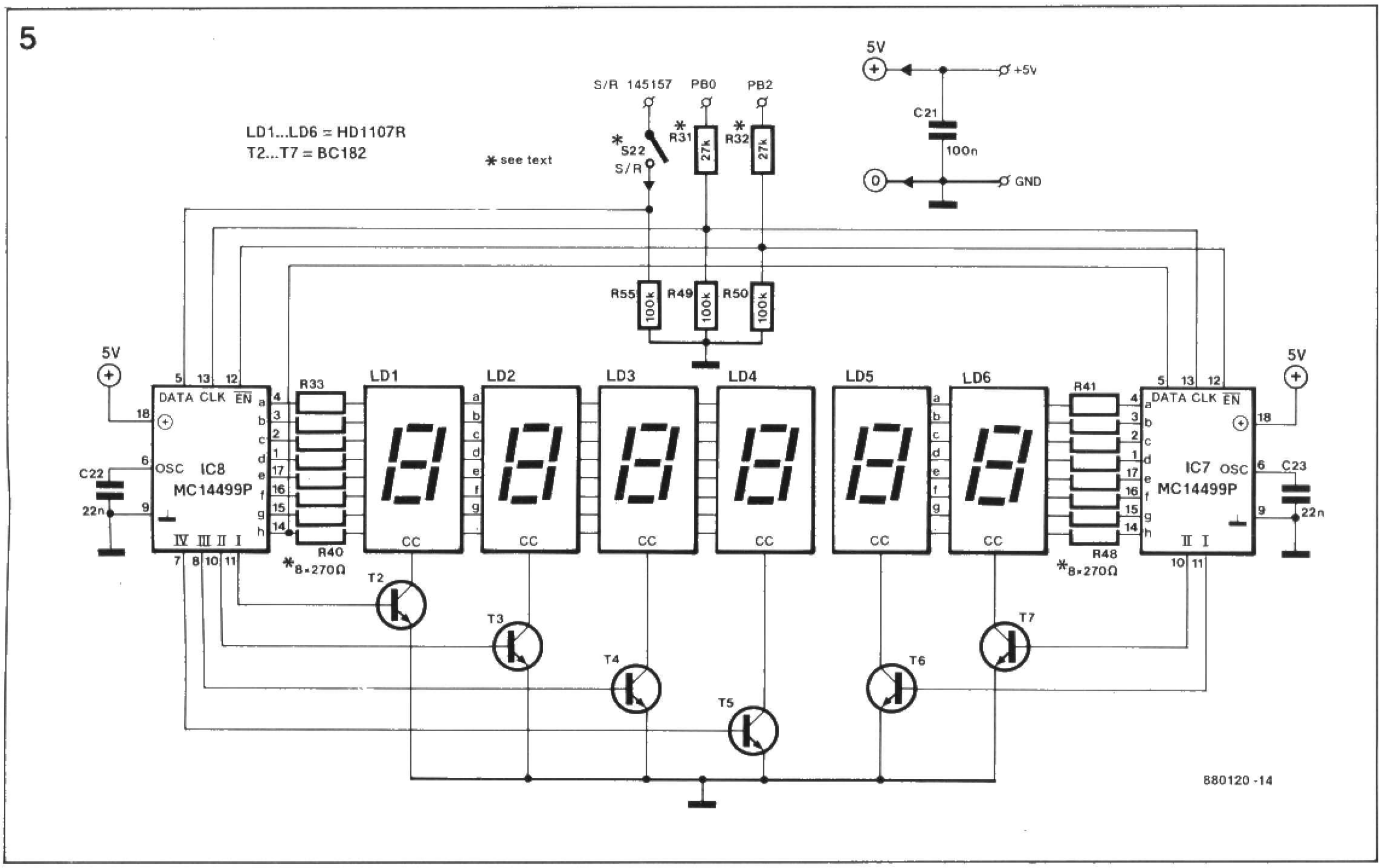


Fig. 5. Six-digit multiplexed LED display.

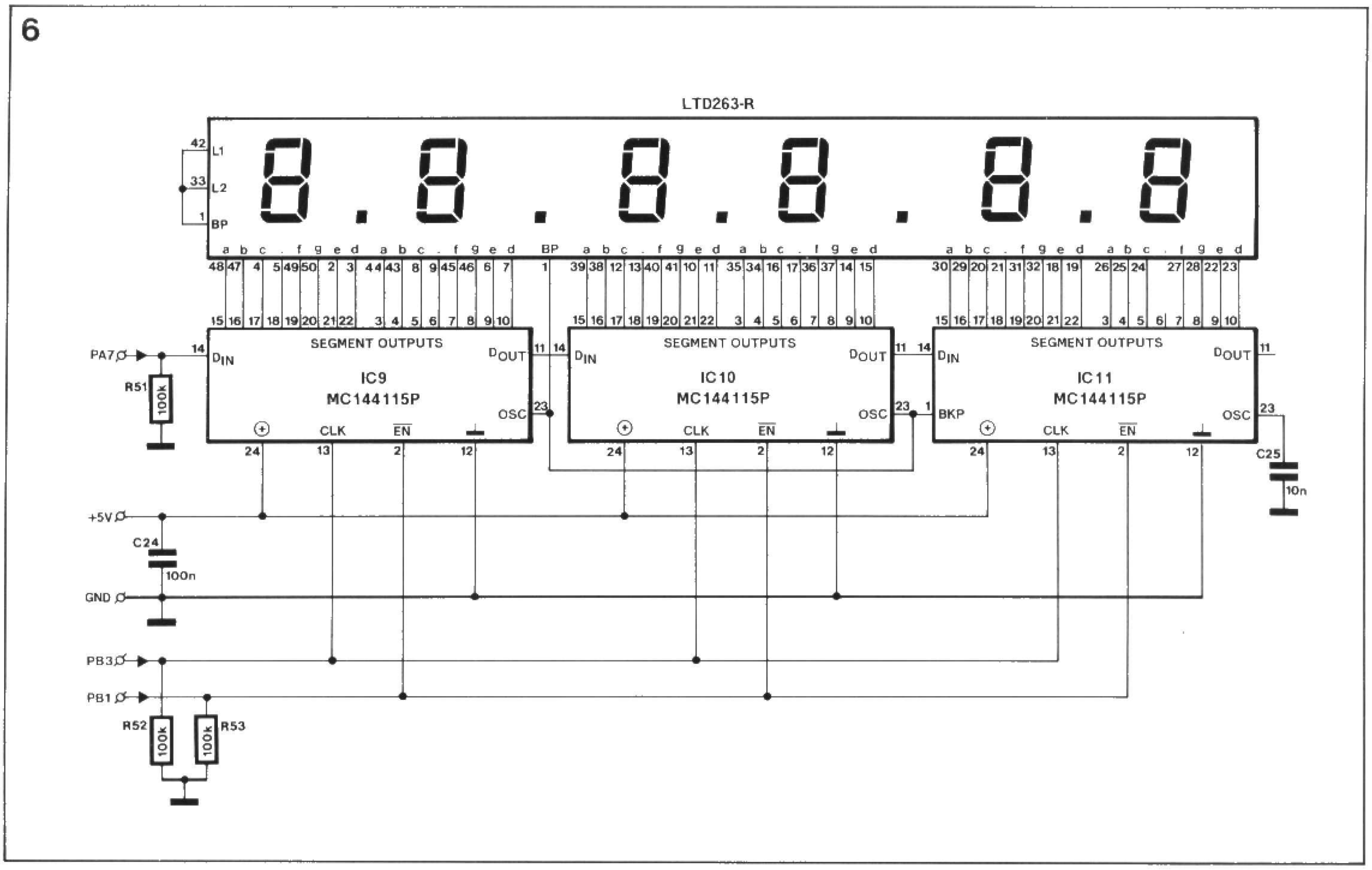


Fig. 6. Static liquid crystal display for the synthesizer.

display): the current frequency has not been written into or recalled from memory.

A choice of two modes permits the minimum number of keystrokes regardless of method of use. In the station mode, previously memorized stations can be recalled by pressing only the required key — there is no RECALL key. Storing a frequency requires two key actions, viz. STORE (indicated by a decimal point on the left-most digit) followed by the memory number.

If direct frequency entry is required, MODE is pressed followed by the frequency. There is now a choice: press MODE again to jump to the new frequency, and return to the station mode. Alternatively, press EXECUTE to jump to the selected frequency but stay in the frequency mode — new frequencies can then be selected with only the EXECUTE key required after each new frequency entered. The store facility also works in the frequency mode.

If it is desired to change back from frequency mode to station mode without retuning the radio, press STORE, then MODE, and to display current frequency press EXECUTE. In station mode, EXECUTE updates the synthesizer and the display with the current frequency. This can be used when the receiver is newly switched on to retune the frequency which was in use when it was switched off, even if that frequency was not stored. Returning to the circuit diagram of Fig. 3, it is seen that this is achieved by T_1 resetting the microprocessor when the radio is switched on — when the 10 V supply rises, T_1 momentarily pulls the CPU's reset input low.

Bands and IF offsets

Port B lines PB4...PB7 incl. are used for providing band selection information to the CPU. These lines can be tied to the appropriate logic level if only one band is required (one band can constitute all the bands which use the same oscillator, but select frequency range by switching inductors). If, however, more than one oscillator has to be tuned, or the step size has to be changed (e.g. between MW and SW), the port lines can be set to the appropriate logic level by means of a set of switches (as shown in

Table 1

Band	PB7 (S18)	PB6 (S19)	PB5 (S20)	PB4 (S21)	IF offset (kHz)	Step (kHz)	Memory	Use	Prescaler
0	0	0	0	0	455	5/1	1	SW	–
1	0	0	0	1	468	5/1	1	SW	–
2	0	0	1	0	470	5/1	1	SW	–
3	0	0	1	1	10,700	5	1	SW	÷ 5
4	0	1	0	0	– 10,700	50/10	2	FM	÷ 10
5	0	1	0	1	0	50/10	2	FM	÷ 10
6	0	1	1	0	– 70	50/10	2	FM	÷ 10
7	0	1	1	1	10,700	50/10	2	FM	÷ 10
8	1	0	0	0	455	9/1	3	MW	–
9	1	0	0	1	468	9/1	3	MW	–
10	1	0	1	0	470	9/1	3	MW	–
11	1	0	1	1	10,700	5	3	SW	÷5

Table 2

```
0000:  2E 01 80 CD 18 4D 24 23 5F B7 28 A6 27 B7 2D A6
0010:  0E B7 2E D6 18 71 B1 28 27 0A A1 77 27 0D 5C 5C
0020:  5C 5C 20 EF 5C DD 18 71 CD 1A E2 80 A6 F0 B7 04
0030:  A6 0F B7 05 3F 01 3F 00 3F 2F A6 27 B7 2D A6 0E
0040:  B7 2E CD 19 63 10 2F CD 1A E2 8E 20 FD A6 F7 48
0050:  24 12 B7 00 2F F9 98 B6 00 AD 0C 2F 07 2E FE AD
0060:  06 2E FA 99 3F 00 81 AE FF 21 FE 21 FE 5A 26 F9
0070:  81 EE CC 18 B1 DE CC 18 B1 DD CC 18 B1 DB CC 18
0080:  B1 BE CC 18 B1 BD CC 18 B1 BB CC 18 B1 7E CC 18
0090:  B1 7D CC 18 B1 7B CC 18 B1 ED CC 19 91 D7 CC 19
00A0:  3D E7 CC 19 12 EB CC 19 A2 B7 16 2F 81 77 CC 19
00B0:  05 9F 44 44 00 2F 1E 06 2F 1B B7 24 05 2F 05 15
00C0:  2F CD 1A C2 CD 1A D9 BE 22 E6 01 F7 5C B3 23 26
00D0:  F8 B6 24 F7 81 97 B7 2E D6 1A CF B7 2D 9F 4C 0F
00E0:  01 04 AB 14 20 05 0D 01 02 AB 0A 48 97 07 2F 0B
00F0:  B6 30 E7 30 B6 31 E7 31 17 2F 81 E6 30 B7 30 E6
0100:  31 B7 31 20 5E AD 1A 3C 30 26 02 3C 31 5A 26 F7
0110:  20 51 AD 0D 3D 30 26 02 3A 31 3A 30 5A 26 F5 20
0120:  42 AE 02 02 2F 0D AD 0C A1 03 27 07 AE 0A 0F 01
0130:  02 AE 12 81 B6 01 A4 70 44 44 44 44 81 00 2F 23
0140:  CD 19 B6 AE 10 BF 2B CD 1A 5B AD E8 A1 03 26 10
0150:  AE 10 BF 2C CD 1A 5B AE 05 E6 0F E7 10 5A 26 F9
0160:  CD 1A 0B CD 1A 91 AD CC A1 03 26 19 AE 10 BF 2B
0170:  BF 2C AE 16 CD 1A 5B AE 16 BF 2B AE 10 CD 1A 5B
0180:  AE 10 CD 1A 5B CD 19 B6 AE 10 BF 2B 14 2F CC 1A
0190:  42 06 2F 02 AD A7 17 2F 01 2F 04 11 2F 20 06 10
01A0:  2F 81 00 2F 05 CD 1A C2 20 09 02 2F 04 12 2F 20
01B0:  02 13 2F 17 2F 81 CD 19 34 48 B7 22 48 BB 22 AB
01C0:  05 B7 23 A6 06 B7 2A BE 23 D6 19 DB 3A 23 BE 2A
01D0:  E7 15 3A 2A 26 F1 AE 16 BF 2C 81 00 00 00 04 05
01E0:  05 00 00 00 04 06 08 00 00 00 04 07 00 00 01 00
01F0:  07 00 00 09 09 08 09 03 00 00 00 00 00 00 00 09
0200:  09 09 09 09 03 00 00 01 00 07 00 3F 30 3F 31 5F
0210:  B6 30 48 B7 22 39 31 B6 31 B7 23 B6 22 48 39 31
0220:  48 39 31 BB 22 B7 30 B6 31 B9 23 B7 31 5C E6 10
0230:  BB 30 B7 30 4F B9 31 B7 31 A3 05 26 D3 38 30 39
0240:  31 81 BF 26 BE 2C A6 06 B7 2A A6 09 E0 05 E7 05
0250:  5A 3A 2A 26 F5 3F 29 3C 29 20 04 3F 29 BF 26 A6
0260:  06 B7 2A BE 2B BF 24 BE 2C BF 25 BE 24 E6 05 3A
0270:  24 BE 25 EB 05 3A 25 BB 29 3F 29 AD 0F BE 26 E7
0280:  05 3A 26 3A 2A 26 E4 81 A0 0A 3C 29 A1 0A 24 F8
0290:  81 B6 31 B7 23 B6 30 B7 22 AE 1C BF 2B AD 25 3C
02A0:  21 AD 1F A6 0E B7 27 34 23 36 22 34 23 36 22 24
02B0:  06 AE 10 BF 2C AD A4 AE 1C BF 2C AD 9E 3A 27 26
02C0:  EA 81 AE 10 A6 06 B7 2A 7F 5C 3A 2A 26 FA 81 EB
02D0:  60 C7 E5 6C AD AF E0 EF ED A6 10 B7 22 AB 05 B7
02E0:  23 81 AD F5 AE 1C AD DC BE 22 5A 5C B3 23 27 03
02F0:  F6 27 F8 5A BF 27 A6 05 B7 25 BE 23 F6 BF 26 97
0300:  D6 1A CF BE 25 E7 1C 3A 25 BE 26 5A B3 27 26 EC
0310:  0D 01 05 0E 01 02 18 1F B6 1C 01 2F 02 B6 2D 07
0320:  2F 02 AA 10 B7 1C CD 19 34 A1 03 27 05 03 2F 02
0330:  18 20 AE 05 E6 1C BF 24 1F 00 AE 08 44 24 02 1E
0340:  00 16 01 17 01 1F 00 5A 26 F2 BE 24 5A 2A E5 A6
0350:  FF B7 1F B7 20 B7 21 A6 0F B7 1D 3F 1F BE 23 F6
0360:  B7 22 CD 1C 05 B7 23 A3 15 26 04 B6 23 B7 1E A3
0370:  14 26 04 B6 22 B7 1D B6 21 A3 13 26 06 A4 F0 BB
0380:  22 B7 21 A3 12 26 06 A4 0F BB 23 B7 21 B6 20 A3
0390:  11 26 06 A4 F0 BB 22 B7 20 A3 10 26 06 A4 0F BB
03A0:  23 B7 20 5A B3 27 26 B7 0D 01 05 0E 01 02 10 1F
03B0:  01 2F 0E B6 20 A4 0F B7 20 B6 2E AD 48 BB 20 B7
03C0:  20 07 2F 02 16 1F CD 19 34 A1 03 27 05 03 2F 02
03D0:  1E 1D A6 FF AD 39 A6 FF AD 35 A6 FF AD 31 B6 31
03E0:  AD 28 B6 30 AD 29 14 01 15 01 5F E6 1D BF 24 AD
03F0:  1E BE 24 5C A3 05 25 F3 A6 4E AD 0E A6 21 AD 0F
0400:  14 01 3F 01 81 48 48 48 48 81 48 AE 07 20 02 AE
0410:  08 48 24 02 12 01 11 01 10 01 13 01 5A 26 F2 81
0420:  FF FF FF FF FF FF FF FF FF FF FF FF FF FF FF FF


07F0:  FF FF FF FF FF FF 18 2C 18 2C 18 00 18 2C 18 2C
0800:  00
```

880120

Table 2. Hexadecimal dump of the EPROM contents. EPROM address range 0430_H to $07EF_H$ is purposely not shown because it is left unprogrammed (addresses read FF_H), just as the remaining 6 Kbyte in the EPROM (0800_H—$1FFF_H$).

the circuit diagram), or, if available, spare contacts on the band selection switch in the receiver. Additionally, the local oscillator feed to the MC145157 synthesizer may need to be switched with the aid of RF relays or PIN diodes. The direct tuning voltage will not normally need to be switched as it can be fed to all varicaps in parallel.

The relationship between the bit combinations on the four MS lines of Port B and the selected band plus IF offset is shown in Table 1.

- Bands 0, 1 and 2: single-conversion SW receivers;
- Band 3: dual-conversion SW receivers (external prescaler $P=5$);
- Band 4: 'oscillator-low' FM receivers or front ends such as the LP1186 (external prescaler $P=10$);
- Band 5: FM band without IF offset; intended for seeing the actual oscillator frequency on display in experimental configurations using a purpose-built test oscillator (external prescaler $P=10$; no prescaler in band 13);
- Band 6: low IF (70 kHz) FM radio ICs (e.g. TDA7000);
- Band 7: 'oscillator-high' FM receivers or front ends such as the Toko 5803/4 or 5402, or Larsholt 8319 or 7254;
- Bands 8, 9 and 10: single-conversion MW receivers (9 kHz steps);
- Band 11: dual-conversion SW receivers (external prescaler $P=5$).

It is seen that bit combination 011_2 in PB6, PB5 and PB4 will select 10.7 MHz IF shortwave regardless of the state of line PB7, so two banks (1 and 3) of memory giving a total of 20 stations can be used, provided that the third bank is not being used for medium wave. A front-panel button connected to PB7 is required to utilize this feature. This will also work for the low-IF shortwave options in which the raising of PB7 selects medium wave with the same IF offset.
The IF offsets can be modified in EPROM if required. They are in 6-digit unpacked BCD format, starting at memory address $1E05_H$ with negative offsets appearing in 10-complement form. FM offsets are in tens of kHz, all others in kHz. For medium wave, starting at band 8, the same series of offsets is used again starting at band 0's 455 kHz. Only the first three are meaningful for medium wave, and at band 11 the software automatically repeats a selection of band 3 as described above. Beyond this there are no useful bands except, perhaps, band 13 which, like band 5, has a zero IF offset.
The software does not include any restrictions on the frequencies which can be used in each band. This maximizes the versatility of the synthesizer. For example, the shortwave bands can be used for MW in the USA where the

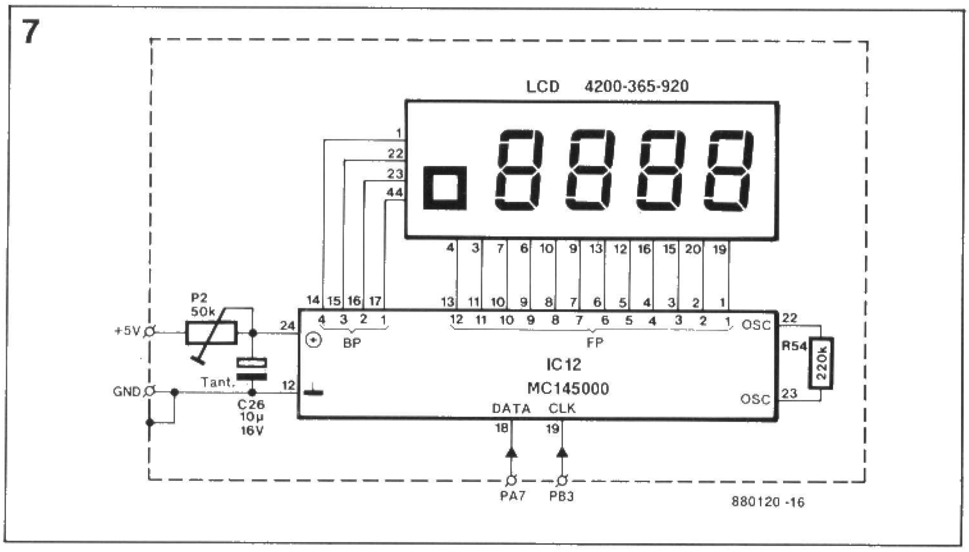


Fig. 7. Multiplexed liquid crystal display. P2 is used for setting the contrast.

channel spacing of 10 kHz means that the SW step size of 5 kHz is more useful than the 9 kHz provided on MW for use in Europe.

Select your display

The display indicates the current frequency and memory number, and assists with the entry of commands and new frequencies. The user is offered a choice of three types of 6-digit display for the synthesizer:

1. LED display; this has the advantage of probably being the least expensive thanks to the use of common 7-segment LED displays. The disadvantage of this circuit is its relatively high current consumption.
2. Static liquid crystal display.
3. Multiplexed liquid crystal display.

All three displays are driven with only two or three lines from the processor board to simplify wiring, and they work without any change in the software. It is, of course, possible to omit the display altogether, or use two displays simultaneously.

The multiplexed LCD is definitely the most elegant of the three options available, since it enables building a compact display unit with only one driver chip and few connections between this and the LCD. Unfortunately, however, the display required in this application proved very difficult to obtain, and it was, therefore, decided not to support this option with a printed-circuit board. The circuit diagram will, however, be discussed below.

LED display

The circuit diagram is given in Fig. 5. In line with the rest of the design, the LED display driver is also based on low-power CMOS LSI chips from Motorola, in this case two Type MC14499 display decoders/drivers set up in a multiplexed circuit with four and two common-cathode 7-segment LED displays. The anode resistors, R_{33}...R_{40} incl. and R_{41}....R_{48} incl. should be dimensioned to give the required compromise between brightness and power consumption. IC_7 and IC_8 share their clock and latch enable (LE) lines with the synthesizer (IC_1), and receive their data from its shift register (S/R) output, pin 12.

As the current consumption of the LED display unit is of the order of 100 mA using 270 Ω anode resistors, the module cannot be left on with the microprocessor in battery-backed up applications, but should be switched off with the receiver. As the data to the drivers is supplied by the MC145157, the display should not be switched off while the MC145157 is still powered, unless the data line (SR) from this is also disconnected by opening S_{22}.

Static LC display

6-digit static displays are currently available with standard pin-outs from a number of manufacturers, with only minor differences in the use of colons and other signs, which are not used here. Figure 6 shows the circuit diagram of the static LC display unit set up around drivers Type MC144115P. The suggested display from Philips Components (formerly Mullard/Videlec) gives good contrast while requiring very little power — the total current consumption of this display module is about 50 µA. Non-used segments in the display are tied to the backplane.

Multiplexed LC display

The software can control a display unit composed of a 6-digit multiplexed LCD and a single driver chip as shown in Fig. 7. The benefits of a multi-plane display are immediately apparent from a comparison of the number lines between the controller and the display in this circuit diagram and that of Fig. 6. The controller, a Type MC145000, is fed serially with 48 bits corresponding to 6 digits of 8 segments, including the decimal point. It formats this data into the four backplane and 12 front-plane waveforms required to drive the LCD. Preset P_2 is used for setting the contrast. To avoid interference with the radio owing to multiplexing pulses, the complete display module should be fitted in a metal enclosure.

Part 2 of this article will deal with the prescalers and the construction of the synthesizer.

COMPONENT NEWS

Mullard no more

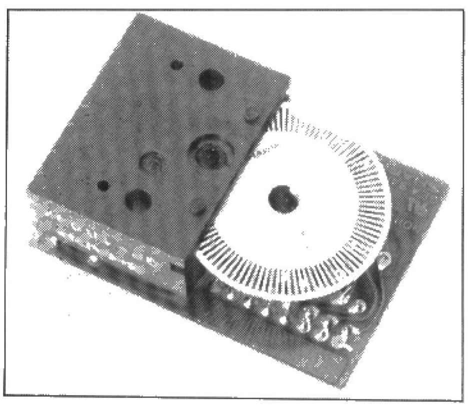
After 68 years in the forefront of Britain's electronics industry, 'Mullard' has ceased to exist, at least the name. The company, acquired in 1927 by the Dutch Philips Group, is now trading under the name 'Philips Components'. The new name has also been adopted by Philips's world-wide component operation, formerly known as 'Elcoma'.

Optical shaft encoder

An optical shaft encoder, made by Sharp, is now available from Greenweld at a very competitive price.

Producing two phase-shifted outputs and a sync pulse once every revolution, it is ideal for use in robotics applications or, in fact, anywhere where the position and speed of a rotating shaft need to be known.

The encoder costs £8.50 incl. VAT and is supplied with comprehensive data sheet. See p. 10 for Greenweld's address.



DIRECT-CONVERSION RECEIVER FOR 80 METRES

What better occasion to present an ultra-simple 80 m receiver than this month's issue devoted to amateur radio and TV? An ideal project for the holidays, this CW/RTTY/SSB receiver is inexpensive, yet has good sensitivity and selectivity. You will have it ready in no time, and it only requires headphones, a set of batteries and a long-wire aerial to bring in the 80 metres band, popular among hams around the world for its reliable propagation characteristics.

The tuning range of the receiver discussed here is about 3.5 to 4.0 MHz, a section of which is assigned to licenced radio amateurs. In most areas in the world, the 80 m amateur band extends from 3.5 MHz to 3.8 MHz. Predominant modulation methods are single-sideband (SSB), continuous wave (CW) and FSK RTTY (radioteletype based on frequency shift keying, using an SSB transmitter). The 80 m band has some interesting properties as regards propagation. Daytime range is usually of the order of a few hundred kilometres, while in the evening and at night field-strength increases, and stations up to 2,000 kilometres away can be heard. Occasionally, American stations are received in Europe in the early hours of the morning.

Direct-conversion receiver

The operating principle of the direct-conversion (or homodyne) receiver is simple: the received signal is mixed with that of a local oscillator to give a beat note, which is at once the AF output (see Fig. 1). In a direct-conversion receiver,

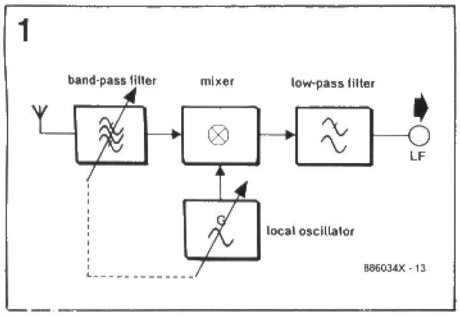


Fig. 1. Basic direct-conversion receiver.

the degree of selectivity is determined almost exclusively by the AF low-pass filter. The present design uses an active mixer which doubles as a linear detector, obviating the need for a prestage and providing the necessary high audio amplification thanks to a special feedback circuit.

The practical circuit of the receiver is shown in Fig. 2. The circuit looks relatively complex at first, but is basically not very different from the block diagram.

The aerial signal is fed via coupling capacitor C_1, to a parallel tuned circuit, L_1-C_2-C_3-C_4-D_1. This band-pass filter can be tuned by applying a direct voltage to variable capacitance diode (varicap) D_1. Coupling capacitor C_5 feeds signals passed by the filter to operational transconductance amplifier (OTA) IC_1, a Type CA3080. An OTA differs from an operational amplifier in supplying an output current rather than an output voltage. In the present application, amplification of the CA3080 is controlled by the feedback circuit and the voltage on pin 5, which is provided by the local oscillator, to achieve the desired mixing effect. The feedback network is composed of C_8, C_{10}, C_{11}, T_1, R_8, R_6, R_4 and C_6. It is a relatively complex circuit because it functions as a current-to-voltage converter in conjunction with FET T_1, and at the same time as a low-pass filter whose roll-off frequency is determined mainly by the capacitance across R_8, i.e., C_{11} (100nF; CW/RTTY) or $C_{11}+C_{12}$ (570nF; phone).

The oscillator set up around T_2 is a varactor-tuned Clapp type with polystyrene capacitors to ensure opti-

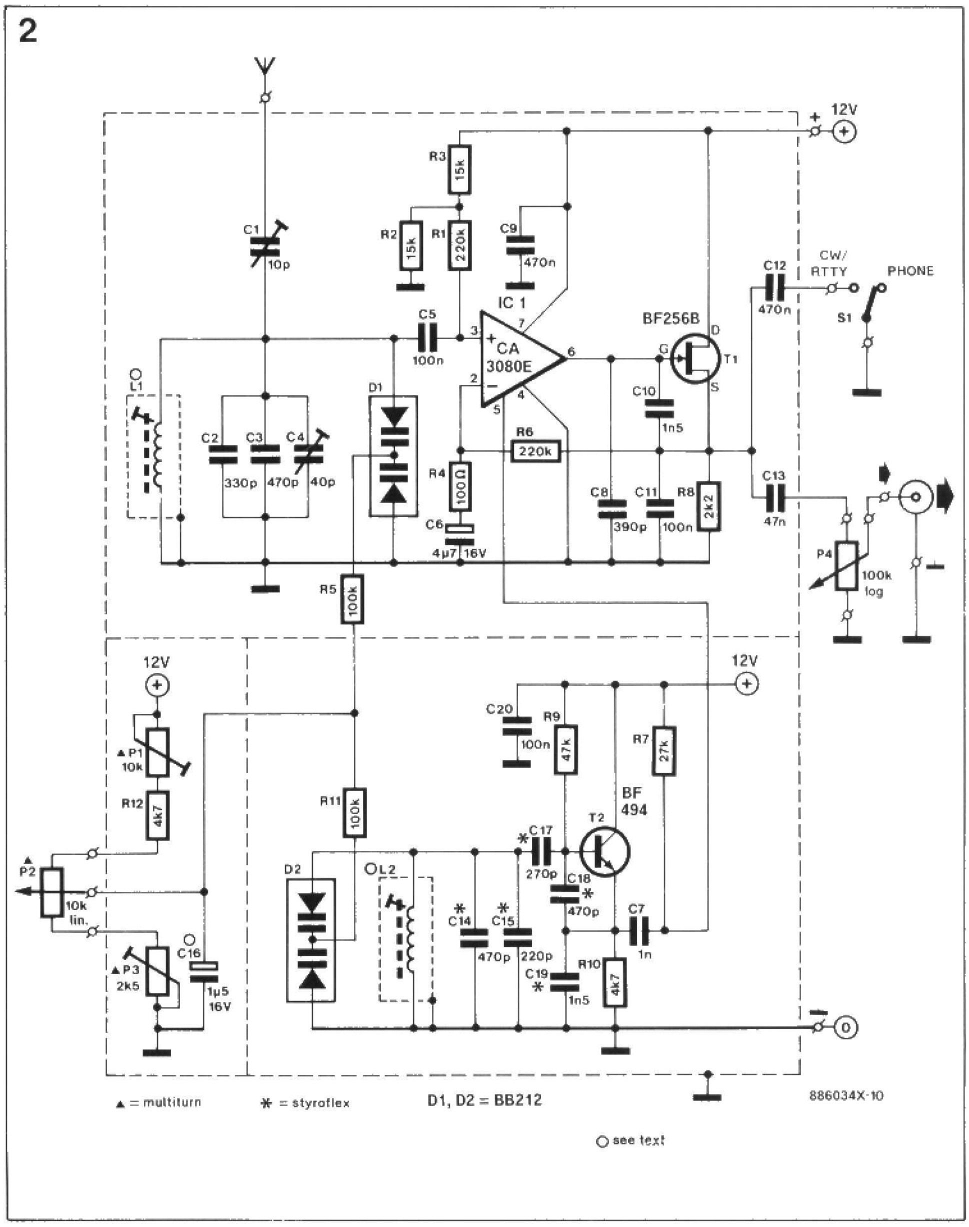


Fig. 2. Circuit diagram of the direct-conversion receiver for the 80 metres band.

3

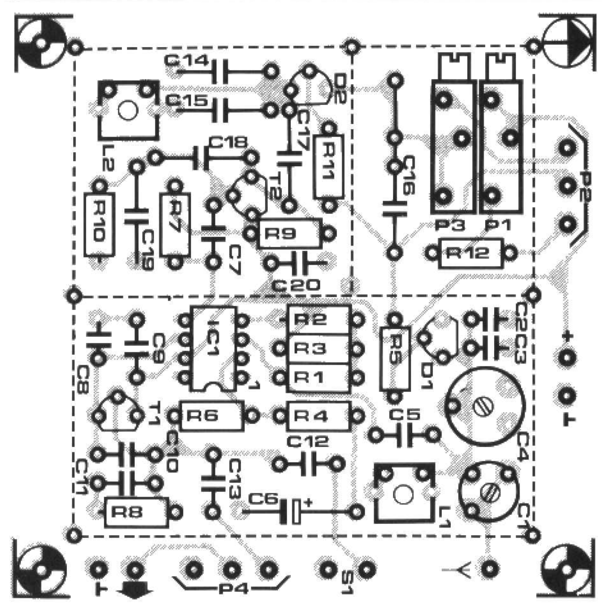


Fig. 3. Track layout and component mounting plan of the double-sided printed circuit board for the direct-conversion receiver.

Parts list

Resistors (±5%):
R_1;R_6 = 220K
R_2;R_3 = 15K
R_4 = 100R
R_5;R_{11} = 100K
R_7 = 27K
R_8 = 2K2
R_9 = 47K
R_{10};R_{12} = 4K7
P_1 = 10K multiturn preset
P_2 = 10K linear multiturn potentiometer
P_3 = 10K multiturn preset
P_4 = 100K logarithmic potentiometer

Capacitors:
C_1 = 10p trimmer
C_2 = 330p
C_3 = 470p
C_4 = 40p trimmer
C_5;C_{11};C_{20} = 100n
C_6 = 4µ7; 16 V
C_7 = 1n0
C_8 = 390p
C_9;C_{12} = 470n
C_{10} = 1n5
C_{13} = 47n
C_{14};C_{18} = 470p $^{+}$
C_{15} = 220p $^{+}$
C_{16} = 1µ5 MKT
C_{17} = 270p $^{+}$
C_{19} = 1n5 $^{+}$

$^{+}$ polystyrene (Siemens: styroflex®) capacitor 5% (available from Cricklewood Electronics).

Inductors:
L_1;L_2 = Neosid assembly 7A1S; winding details are given in the text. Neosid part no. 06955500. Neosid Limited • Icknield Way West • LETCHWORTH SG6 4AS. Telephone: (0462) 481000. Telex: 826405. Fax: (0462) 481008 (contact Mr. E. Adcott). Neosid inductor assemblies are also available from C-I Electronics • P.O. Box 22089 • 6360 AB Nuth • The Netherlands.

Semiconductors:
D_1;D_2 = BB212 (Cirkit stock no. 12-02045)
T_1 = BF256B
T_2 = BF494
IC_1 = CA3080E

Miscellaneous:
S_1 = miniature SPST switch.
PCB Type 886034X (see Readers Services page).

mum stability. The common tuning voltage for the oscillator and the input bandfilter is obtained from multiturn potentiometer P_2. Presets P_1 and P_3 form part of a potential divider, and enable accurate setting of the receiver's tuning range. Since the capacitance of a varicap is inversely proportional to the reverse voltage across it, the received frequency increases with the voltage on the wiper of P_2. With this in mind, it is readily seen that P_1 and P_3 set the upper and lower band limit respectively.

Construction

The direct-conversion receiver is constructed on the double-sided, but not through-plated, printed circuit board shown in Fig. 3. The component side of this board is left largely unetched to enable it to function as a ground plane.

Commence the construction with soldering 15 mm high tin plate or brass screens onto the component side — the locations are shown in dashed lines on the overlay. Use soldering pins at the corners to aid in positioning and, if necessary, bending the screens. Drill holes to ensure access to the spindles of multiturn presets P_1 and P_3 later.

Inductors L_1 and L_2 are home-made and identical. They are composed of 20 turns of 0.2 mm dia. enamelled copper wire wound on the plastic former of a Type 7A1S inductor assembly from Neosid. Remove the three pins on one side of the base, and use the remaining two pins for connecting the winding. Scratch off the enamel coating with a pen-knife, pretin the wire end, remove solder flux by scratching again, and then wind the wire end two times around the pin. Tighten the winding and move it up towards the base to prevent a short-circuit with the ground plane. Solder fast to prevent the base melting and the pin being dislocated. Now close-wind 20 turns of the wire upwards on to the former, right up to the rim. Adjustment of the inductor is facilitated when the grounded ('cold') end of the inductor is near the rim on the former. Secure the winding with a few drops of glue or wax, and check continuity at the pins.

Carefully mount the inductor on the board, and solder the pins at the track side. Fit the ferrite cup, and insert the core. Finally, mount the screening can, and solder the tabs to ground *at both sides of the PCB*.

The mounting of the remaining components on the board is a matter of routine. One terminal of the following components is soldered to ground at both sides of the PCB: R_2, R_8, R_{10}, C_2, C_3, C_6, C_8, C_{14}, C_{15}, C_{19}, C_{20}, D_1, D_2, P_4 and S_1. The two rotor terminals of foil trimmer C_4, and the two ground pins (supply and AF output), are soldered likewise. In the case of the trimmer, take care not to damage the PTFE material by overheating it with the iron.

Unfortunately, there is a small error on the ready-made PCB for this project: facing the flat side of varicap D_1 (BB212), the right-hand side terminal of the device should not be inserted in the hole provided. Instead, it is soldered *direct to the ground surface*. PCBs supplied through our Readers Services are accompanied by a note advising of this design error.

Capacitor C_{16} is preferably an MKT type. When this is not available, a low-leakage electrolytic type may be used instead. The photograph of Fig. 4 shows a prototype of the receiver. Note that a number of ceramic capacitors are fitted in positions that should have polystyrene types. Unfortunately, these were not available when the photograph was taken. The previously mentioned grounded terminal of D_1 is clearly visible to the left of the device. It is essential that the receiver is fitted in a metal enclosure. Figure 5 shows a suggested front-panel layout.

Setting up

The following items are required for set-

5

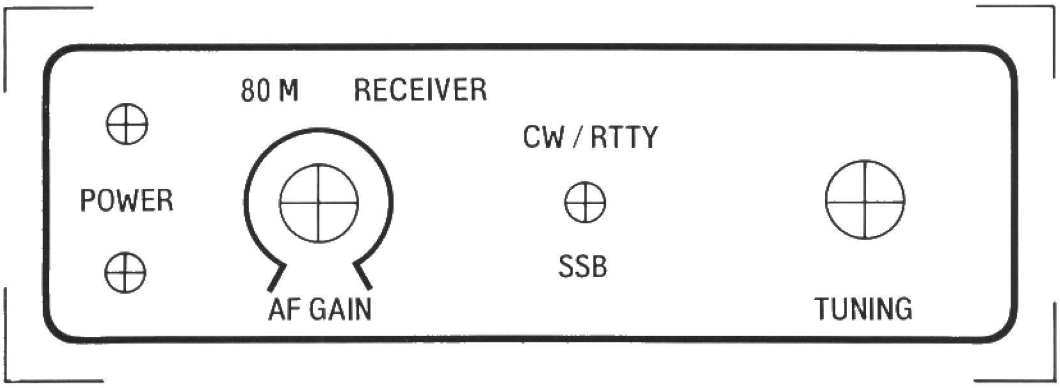
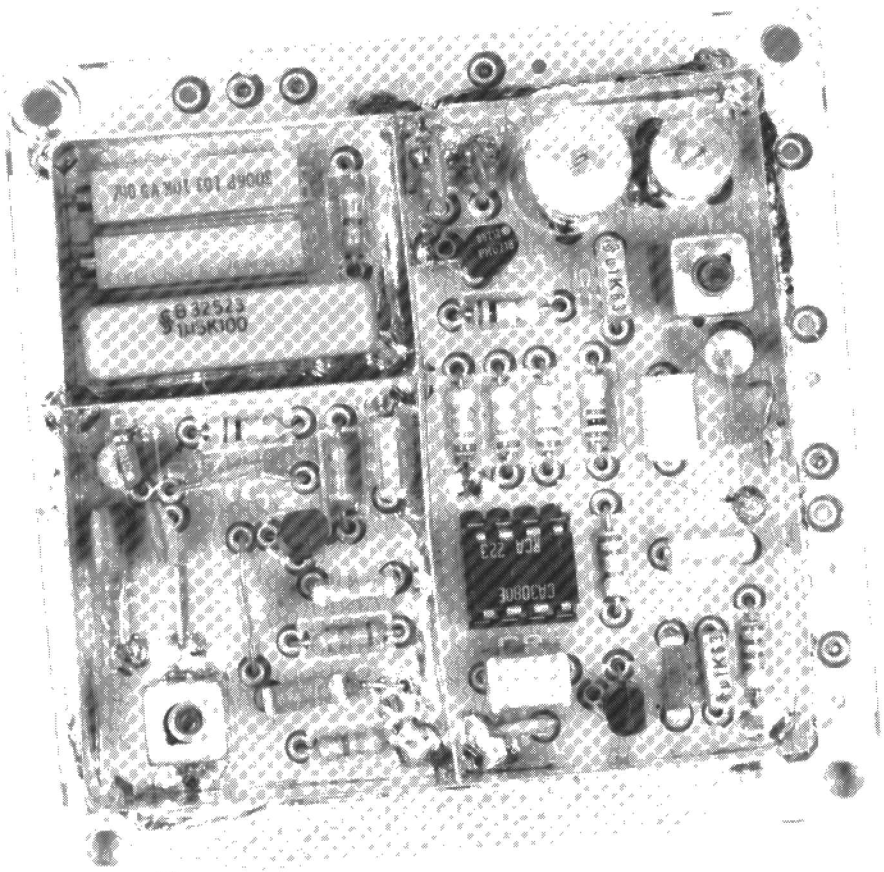


Fig. 5. Suggested layout of the front panel.



Prototype of the receiver board.

ting up the receiver: a nylon trimming tool, a multimeter and a frequency meter (or a good-quality 80 m receiver).

Temporarily power the completed receiver from a regulated 12 V supply. Connect the AF output to an amplifier. Turn P_2 to the centre of its travel, and measure the voltage at the wiper. Adjust P_1 and P_3 to obtain 4 V. Connect the frequency meter to the emitter of T_2, and use the nylon trimming tool to adjust the core in L_2 until the oscillator produces 3.65 MHz. Turn P_2 fully counter-clockwise, and check that the oscillator frequency decreases. Exchange the wires to the outer connections of the potentiometer if the frequency increases.

Adjust P_3 until the frequency meter reads 3.4 MHz. Turn P_2 fully clockwise, and adjust P_1 for an oscillator frequency of 3.9 MHz. Turn P_2 to the centre of its travel, and check that the oscillator frequency is about 3.4 MHz. If necessary, re-do the adjustments of P_1 and P_3 until P_2 covers the full tuning range of 3.4 MHz to 3.9 MHz.

As a matter of course, the function of the frequency meter can be taken over by a calibrated 80 m receiver, which should have no difficulty picking up stray radiation from T_2. The oscillator frequency can then be read from the dial of the auxiliary receiver.

Connect the aerial to the input of the direct conversion receiver. Tune to a weak transmission at the lower band edge (3.4 MHz), and peak C_4 for optimum reception. Then tune to a station at about 3.9 MHz and similarly adjust the core in L_1. Repeat the adjustment of C_4 and L_1 to optimize sensitivity across the band.

Power supply: beware of adaptors

The receiver is preferably fed from a well-regulated 12 V supply, or a set of batteries that gives the same voltage. In many cases, however, it will be convenient to power the receiver from an available 12 VDC mains adaptor. When this is used, it is likely to cause hum in the receiver, a problem which can often be solved by fitting an additional smoothing capacitor of 470 or 1000 μF across the adaptor's output terminals. If hum persists, it is probably caused by capacitive coupling of the receiver and the mains. When the rectifier diodes in the adaptor are reverse-biased, they behave as capacitors. This is not normally a problem, but the oscillator signal, via the supply wires, can thus find a path to these diodes, which cause amplitude modulation of 50 (60) Hz (single-phase rectifier) or 100 (120) Hz (double-phase rectifier). This modulated signal is radiated by the mains wires, and is picked up again by the receiver. In this, the oscillator frequency equals the received frequency, so that hum is produced as the AF output signal.

The supply connections of the receiver should be decoupled by fitting chokes between the terminals of the 12 VDC input socket and the relevant soldering pins on the board. Both socket terminals are decoupled to ground with a 100 nF capacitor. The L-C networks prevent the oscillator signal being superimposed onto the supply lines. If hum still persists, try winding the supply wires through a ferrite ring core, or around a ferrite rod. A final tip is to solder a 100 nF capacitor across each rectifier in the mains adapter. *B*

NEW LITERATURE

CIRCUITS, SIGNALS & DEVICES

by Michael Julian
ISBN 0 582 99467 5
515 pages — 238×164 mm
Price £14.95 (soft cover)

Circuits, Signals & Devices is a textbook intended for first- and second-year undergraduates and diploma students in electrical and electronic engineering. It will also be of value to students studying systems and electronics in other engineering and scientific disciplines.

The mathematical content is broadly consistent with the intended readership. The text is written for readers meeting a topic for the first time. To meet the needs of an evolving technology, the book examines not only the more traditional areas of circuits and devices, but also covers important aspects of signal analysis which are often omitted from introductory books.

Consistent with its title, the book is divided into three sections. The first, Circuits, starts with basic principles and progresses through circuit analysis and techniques to time and frequency response.

The second section, Signals, commences with Fourier Series and proceeds through modulation techniques to noise in transmission systems.

Although the first two sections deal with the subject matter in a clear and direct manner, the third section, Devices, is, for me at least, the most fascinating part of the book. It gives a wealth of information on all kinds of semiconductor devices — from materials and bipolar transistors to microprocessor systems.

The book is well illustrated and indexed. Each chapter ends with a useful selection of exercises, answers to which are given at the end of the book.

A clear, useful textbook that will no doubt find its way to many a college or university library and also to students' own bookshelves. The very reasonable price is an additional bonus.

Longman Scientific & Technical
Longman House
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HARLOW CM20 2JE

Mobile Radio Telephones in the UK

by Dr R.C.V. Macario
ISBN 1-85181-182-6
194 pages — 225×155 mm
Price £9.95 (soft cover)

This lavishly illustrated book, written by one of the UK's authorities in the field of mobile radio telephones, provides a clear guide to this topical subject. Dr Macario deals with the subject in a manner that will enable the reader to get to grips with the technical, social, and legislative aspects of mobile radio telephones in an enjoyable way.

The technology is described and placed in an historical context, enabling the reader to appreciate the sources of technical innovation and understand the steps that have led to the present revolution in communications. The geographical aspects of both transmission and reception are covered and the limitations of the available spectrum discussed. The various technologies available for radio telephone usage are explained and future possible developments described.

The regulation and licensing of the radio spectrum are covered and an appendix outlines the present allocation of frequencies.

The text is confined to UK allocations, activities, and developments in mobile radio, unless a common European policy becomes applicable (as, for instance, in cellular radio).

This book is in no way meant to be a textbook, manual, or system guide. Its primary intention is to be a text that can be read easily, say on a train, or late at night, to give added awareness to the remarkable engineering achievements of mobile radio. It should please many of the tens of thousands of radio mobile users who want to know a little more about their equipment and its background.

Glentop Press Ltd
Bath Place
BARNET EN5 5XE

VIDEO TECHNIQUES

by Gordon White, CGIA, CEng, MIERE
ISBN 0-434-92290 0
496 pages — 220×145 mm
Price £30 (hardback)

This second edition of *Video Techniques*, which provides a comprehensive treatment of the many aspects of video for the engineer or technician working in television or associated industries, has been updated throughout and enlarged. Amply illustrated (235 diagrams and 100 half-tones), *Video Techniques* does not describe circuitry or individual pieces of equipment, unless it is to illustrate a principle, as these change rapidly. Rather, it describes the principles of television and shows how the equipment is designed and functions in the complete system and, with present technology, its capability and limitations. These principles generally remain the same irrespective of the model, which usually involves only modifications to the performance or operation.

Contrary to so many other books on video, it does not ignore different transmission systems, such as SECAM, and NTSC. It is right up to date with descriptions of the MAC system, direct broadcast satellite (DBS), and high-definition television (HDTV).

Other contents include standards convertors, CCD cameras and telecine, camera tubes, broadcast video recorders, ENG cameras and recorders, domestic recorders and cameras, video tape, video discs, interactive video, conference television, fibre optics, television receivers and displays including FST, broadcast monitors, teletext, large screen projection, world standards, digital techniques, and graphics and digital effects generators.

A book that nobody involved in whatever aspect of video engineering, production, or management can afford to be without.

Heinemann Professional Publishing
22 Bedford Square
LONDON WC1B 3HH

BRITISH STANDARDS

BS 2011:Part 2.1 Ea:1988
IEC 68-2-27:1987

This revised standard, *Basic environmental testing procedures*, supersedes BS2011: Part 2.1 Ea:1977. It describes the tests applicable to components, equipments, and other electrotechnical products.

BS 6840:Part 12:1987
IEC 268-12:1987

This part of BS 6840, *Sound system equipment: specification for applications of connectors for broadcast and similar use*, supersedes BS 5428:Part 5:Section 5.3.

BS 5310: Part 3A

This additional Part of BS5310, *Specification for hand crimping tools for contacts of electrical connectors*, gives detail requirements for such equipment, incorporating a system of multiple indentors for use with removable male and female contacts complying with BS 9521, electrical connectors complying with BS9210, and components that meet both these standards.

BS 5070:Parts 1 & 2

This standard, *Engineering diagram drawing practice*, has been revised and published in a fresh format. It complements BS 308: *Engineering drawing office practice*, which last year celebrated its diamond jubilee amid considerable professional acclaim.

BS 5783

This revised standard, *Code of practice for handling of electrostatic sensitive devices*, has now been published. It describes practices which, if adopted, will reduce the risk of inadvertent damage to the components. It provides revised definitions and workshop layout drawings and includes revised recommendations.

BS 9530
This is a new standard giving a *Specification for cable fitting accessories of assessed quality for circular electrical connectors: generic data, methods of test, and rules for the preparation of detail specifications.* The specification is intended primarily for use in telecommunication and allied electronic equipment.

BS 6513:Part 5: 1987
BS 6513:Part 6: 1987
These parts of the multipart standard for *Wideband cabled distribution systems* cover (Part 5) *Recommendations for one-way and interactive data services*, and (Part 6) *Specification for safety requirements.*

CECC 52000
This generic specification for the test and measurement procedures of *Mercury wetted change-over reed contact units mechanically biased* has been formally approved by the CENELEC Electronic Components Committee (CECC). CENELEC is the European Committee for Electrotechnical Standardization.

International certification and approval schemes
A third edition is now available of the popular publication *International Certification and Approval Schemes* from BSI Technical Help to Exporters. This edition presents revised data on nine major technical certification schemes which relate to a wide range of products in more than 90 countries.
These schemes, which have been extensively revised (hence the need for a new edition) include the CB scheme for the safety of electrical equipment; the CENELEC Certification Agreement; the CEN Certification System; the CENELEC HAR Agreement: the CECC Harmonized System; the IECQ System; the ECE Harmonization and Conformity Scheme; EEC schemes under 'Article 100' Directives; and schemes coordinated by EFTA.

British Standards may be ordered from
The Sales Department
BSI
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Readers should also note that each county in the UK has at least one large Public Library where complete sets of British Standards are kept for general consultation.

Available free from Hitachi New Media Products is *CD-ROM Directions*, a quarterly journal which reports on the latest developments, standards, software, and available databases for Compact Disk-Read Only Memory optical disk drives. Apply to **Nick Rogers • Hitachi New Media Products • Hitachi House • Station Road • HAYES UB4 3DR • Telephone 01-848 8787.**

Introducing Digital Audio

by Ian R. Sinclair
ISBN 1-870775-05-8
112 pages — 216×138 mm
Price £5.95 (soft cover)
This non-mathematical introduction to CD, DAT, and sampling is intended primarily for students, technicians, and enthusiasts. But, although the mathematical background and theory are omitted, the principles and methods of digital audio are explained in detail. These principles (and practices) owe little or nothing to the traditional linear circuits of the past, and are much more comprehensible to today's computer engineer than the older generation of audio engineers.
Digital recording methods have existed for many years and have become familiar to the professional recording engineer, but the compact disc (CD) was the first device to bring digital audio methods into the home. The next step is the appearance of digital audio tape (DAT) equipment.
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LONDON N3 2AR

Getting the most from your multimeter

by R.A. Penfold
ISBN 0-85934-184-4
102 pages — 178×110 mm
Price £2.95 (paperback)
The multimeter is probably the most useful piece of test equipment for engineers, technicians, students, and enthusiasts alike. In spite of its assumed simplicity, a multimeter is of little use unless you understand how it can be put to effective use, and are aware of its limitations.
The book is aimed at beginners and those of limited experience of electronics. Chapter 1 covers the basics of analogue and digital multimeters, discussing the relative merits and limitations of the two types. In Chapter 2 various methods of component checking are described, including tests for transistors, thyristors, capacitors, and diodes. Circuit testing is covered in Chapter 3, with subjects such as voltage, current, and continuity checks being discussed.
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Shortwave frequency list

by C.J. Broth
ISBN 90-6082-289-7
96 pages — 175×120 mm
Price £4.95 (soft cover)
The easy tuning to, and locating of, transmitters on medium wave and FM is in sharp contrast to the bewildering mass of sounds of all descriptions on the shortwave bands.
This book is intended to help the shortwave listener in identifying the more important shortwave broadcast stations. It lists such stations in order of frequency: from 2,260 kHz to 21,810 kHz. In addition, the lists show the radiated power of most of the stations.
A 'must' for every serious shortwave listener.
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Newnes Audio and Hi-fi Engineer's Pocket Book

by Vivian Capel
ISBN 0-434-90210-1
208 pages — 190×90 mm
Price £9.95 (hardback)
Like its companion volumes (Electronics: reviewed in March 1987; Computer: April 1987; Radio and Electronics: October 1987; Television and Video: November 1987), this latest in the Engineer's Pocket Book series offers a concise collection of practical and relevant data. Topics covered include microphones, gramophones, compact discs, tape recording, high-quality radio, amplifiers, loudspeakers, and public address.
Acoustics is not often dealt with in audio books, nor is it too well understand by many audio engineers. Therefore, a lengthy section on this subject has been included, dealing with most aspects the audio engineer or technician is likely to encounter, from human hearing to sound insulation.
From this ancient art to the modern wizardry of digital recording and the compact disc is an enormous leap that illustrates the wide range of knowledge required of the audio engineer: one that encompasses mechanics, heat, magnetism, semiconductor technology, and electronics.
Newnes Audio and Hi-fi Engineer's Pocket Book is essential reading for sound-system engineers and technicians, hi-fi enthusiasts, dealers, and students.
Heinemann Professional Publishing
22 Bedford Square
LONDON WC1B 3HH

TEST & MEASURING EQUIPMENT

Part 8: Function Generators (2)

by Julian Nolan

Global Type 2002

Global Specialties, a member of the American-owned Interplex Electronics Group, manufacture and retail a large variety of electronic test equipment, including oscilloscopes, power supplies, signal sources, and frequency counters.

The Type 2002 Function Generator is one of a range of signal sources that extends from a portable 150 kHz generator, the Type 555, to a synthesized IEEE programmable generator, the Type 8230.

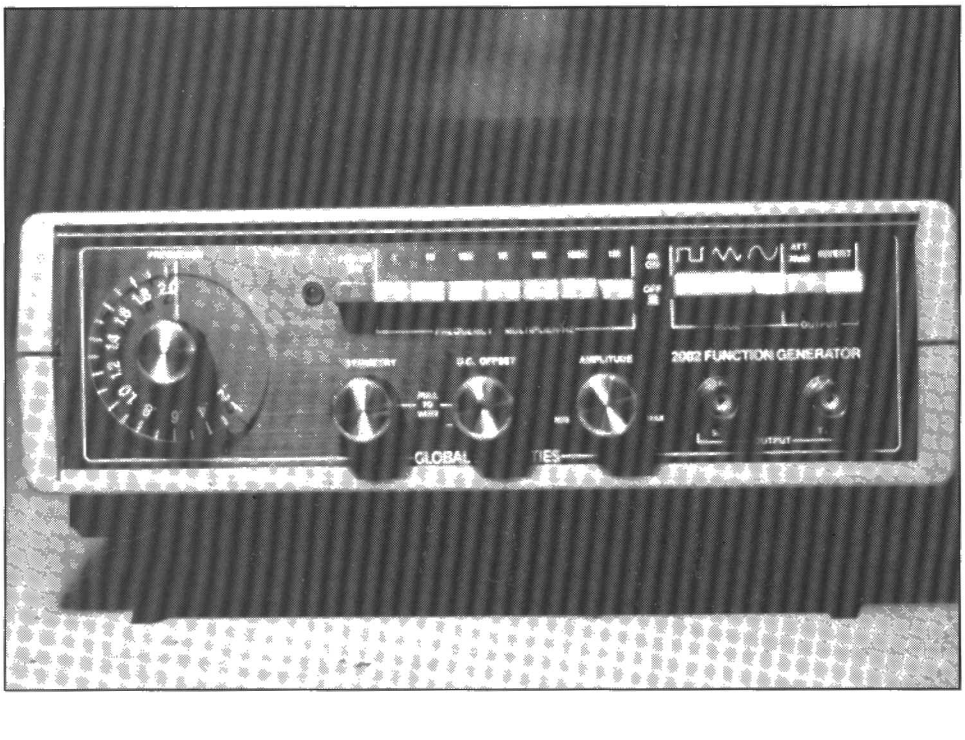
The design of the 2002 is fairly typical in both layout and size. An integral one-position stand enables the 2002 to be operated at an optimum angle when the instrument is not stacked. The stand is of man-made fibre but should be rugged enough for operation in most environments.

The 2002 uses an external a.c. adaptor which is connected to the generator by a 5-pin DIN plug. This somewhat unusual feature would not be an inconvenience if it offered the benefit of (optional) battery operation. The adaptor is set for 240 V a.c. (in the UK) and has no mains voltage selection facility (if operation from a different mains voltage is required, a different adaptor has to be used: this is available on request).

The multiplier ranges and various operating modes are selected by a number of very-closely spaced and somewhat cramped push-button switches. The tight spacing makes operation without pressing more than one switch difficult: the switches may be pressed from the side, but even this is not always successful until a certain measure of familiarization has been achieved.

The output frequency is set by the usual combination of a continuously-variable, logarithmically-calibrated control and seven push-buttons. The specified frequency range is 0.2 Hz to 2 MHz; in practice 0.008 Hz to 2.022 MHz is obtained. Even at the lower frequency, reasonable symmetry and waveform shape are retained. Some frequency instability was apparent at frequencies below 0.2 Hz, but this amounted to nothing much and is in any case to be expected at such low frequencies.

Symmetry is variable over a range of about 1:10 and the invert facility gives added versatility to this. After the warm-up period (about 10 minutes), with a mark-to-space ratio of 1:3 and between the set frequencies of 60 kHz to 190 kHz, latch-up occurs, i.e., the output is a direct voltage at the most negative previous output level. Clearly, this is highly undesirable in any signal source and, although the effect does not occur when the output is inverted, could render the 2002 unusable as a pulse generator between these frequencies.



The general waveform may be selected from sine, triangle, and square, all of which may be varied in terms of symmetry across the range outlined before. In addition, a d.c. output (all mode switches off) is also available and should enable time to be saved in such applications as audio equipment servicing.

Typical preset symmetry is accurately calibrated and within the quoted 1%. Some distortion occurs on both the crest and the trough of a sine wave on all ranges, but this should not affect all but the most critical uses. Apart from this, distortion over all ranges is well within the specified limits.

Level flatness is good with a maximum of variation over all seven ranges not exceeding −0.9 dB.

A −30 dB attenuator is provided to permit low-level signals with a minimum of noise and inconvenience to be obtained. Minimum output in either output mode is effectively at ground level, although a small amount of noise is present. Distortion at the low output levels is very low for a generator of this class.

The rise time of TTL output is good, typically 9 ns (5 ns on the trailing edge). Rise times of the main output were a little higher at 65 ns and 50 ns respectively (these are still better than average).

The VCF input allows a maximum output frequency of 3.2 MHz to be obtained at an input level of −5 V; an input of +10 V gives a sweep range of about 150:1 on all but the lowest ranges. While this may be adequate for a large number of applications, it does not compare well with the range of 1000:1 (or higher) on some other generators in this class.

On the whole, the 2002 is solidly constructed and should be ideal for use in a wide range of operating environments.

External construction is based on a man-made-fibre and aluminium enclosure, which should provide reasonable strength and ruggedness. Why the enclosure is fitted with ventilation vents is not clear.

Internal construction is based on a single PCB which fits snugly in the enclosure. Interconnecting wires between the board and the potentiometers on the front panel and the BNC sockets on the rear panel are kept to a minimum. The board is screened with component identifi-

Table 7

OPERATING RANGE

Frequency range: <0.2 Hz to 2 MHz in 7 calibrated ranges: fine adjustment by vernier, calibrated from 0.2× to 2× main setting.
Frequency accuracy: ±5% of full scale on 1 kHz to 1 MHz; better than ±8% on 10 Hz and 100 Hz ranges.
External sweep range: variable over >100:1 by 10 V_{pp}.
Input impedance: 10 kΩ
Maximum input: ±10 V_{pp}.

OPERATING MODES

Sine wave: distortion <1% from 10 Hz to 200 kHz.
Square wave: mark:space ratio 1:1 ±1% to 100 kHz.
DC range: ±5 V into 50Ω and ±12 V into 1MΩ.
DC off-set: variable ±5 V into 50Ω.

OUTPUTS

50Ω: <0.05 V to 20 V_{pp} from 50Ω ±1% source; <0.025V to 10 V across 50 Ω load; switched attenuator reduces signal and d.c. off-set by 30 dB; output protected against short-circuits.
TTL: capable of driving up to 20 standard TTL loads.

GENERAL
Mains voltage: 110-120-220-240 VAC 50/60 Hz.
Power consumption: 12 VA.
Dimensions: 254×178×76 mm (W×D×H).
Weight: 1.0 kg.
Accessories supplied: mains adaptor; manual.
Warranty: 1 year.

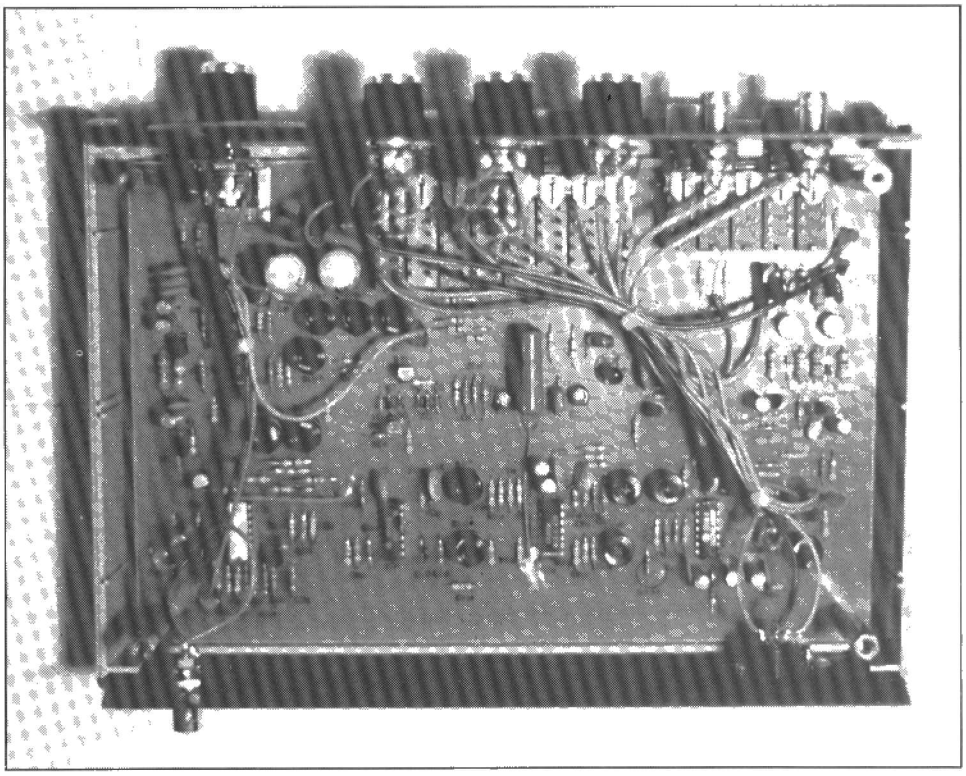


Table 8

	Unsatis-factory	Satis-factory	Good	Very good	Excellent
Dial accuracy				x	
Dial resolution				x	
External sweep range			x		
Distortion			x		
Frequency range					x
Output level range				x	
Internal construction				x	
External construction				x	
Overall specification				x	
Ease of use		x			
Manual			x		

cation numbers, which should be a help when servicing is necessary.

Conclusion

The main strengths of the 2002 are its low-level signal performance and its high grade construction. It also has several poor points, such as the difficulty in operating the range and mode selection switches, and the external power supply. Distortion and frequency range are satisfactory for an instrument in this price range. Moreover, the 2002 provides a good range of functions.
By the time this review is published, an updated version of the 2002 should have become available, which has an improved front panel layout and other features.

The 2002 was supplied by Global Specialties, Unit 1, Shire Hill Industrial Estate, SAFFRON WALDEN CB11 3AQ, telephone (0799) 21682.

Other signal sources available in the Global range

Model 555 – fully portable; 20 Hz to 150 kHz; battery operated; £47.50.

Model 2001 – 1 Hz to 500 kHz frequency range; 20 V_{pp} output; £119.95.

Model 2005 – 50 mHz to 5 MHz sweep-function generator; 1000:1 linear or 10,000:1 log sweep; continuous, triggered and gated modes; distortion 0.5% on 1 kHz and 10 kHz ranges — 1% up to 100 kHz; variable sweep rate; £399.00.

Model 8120 – 1 mHz to 12 MHz AM/FM sweep/function generator; digital frequency read-out; attenuation to 80 dB; continuous, triggered, and gates modes; variable start/stop phase; distortion not greater than 0.5% from 12 Hz to 120 kHz; internal AM (1 kHz); amplifier input (dynamic range ±1 V peak a.c.+d.c.; £895.00.

8200 series – 2 mHz to 20 MHz IEEE programmable function generators; 4 models available: 8210-basic 20 MHz sweep/function generator at £1,895; 8230 — sweep/function generator with PLL counter at £2,295; 8232 — as 8230 but frequency synthesized from 20 Hz to 20 MHz at £2,495; 8241 — pulse/function generator and counter at £2,495.

Levell TG302

Levell have a long history of distributing high-quality test instruments for a large number of manufacturers. The Taiwanese-made TG302 is marketed under the Levell trademark and is only one of a large number of signal generators sold by the company. The TG302 retails at £136.00, excl. VAT.

The generator comes well-packed in a cardboard box, which also contains two probes and the manual. A good length of mains lead is supplied and this is terminated in an IEC plug. The mains voltage setting may be adjusted internally (110, 220, or 240 V a.c.). In contrast to some other instruments in this class, the TG302 is supplied with additional fuses to take account of the different currents that will be drawn at these mains voltages.

The integral 18-position stand conveniently swivels through 360°, enabling the generator to be operated either at an angle or stacked between other instruments.

The appearance of the TG302 is conventional with the function and range selectors mounted along the top of the front panel, with the continuously-variable controls and BNC connectors mounted beneath them.

The output frequency is selected on a logarithmic scale, which covers a ratio of about 50:1. Thus, the output frequency may be varied between 4 kHz and 200 kHz when the 100 kHz range is selected.

The multiplier range is selected by seven push-button switches and is calibrated from 0.02 Hz to 2 MHz. It may sometimes be necessary to increase the duty factor of the signal to bring the output down to 0.02 Hz. Considering the market at which the instrument is aimed, this very slight problem should not deter any users other than those who operate at these low frequencies.

All ranges overlap, so that a continuous range is provided throughout the frequency spectrum covered by the TG302. It should be noted that at the very far end of the scale (fully anticlockwise), the output frequency tends to become unstable.

As already mentioned, the duty factor is variable from a mark-space ratio 1:1 to one about 1:10 or 10:1, depending on the position of the invert switch. These two functions enable a wide range of waveforms to be generated, for instance, a 500 Hz sawtooth with a downward transition time of around 250 microseconds. Pulses can also be generated, although these are rather limited by the maximum mark-space ratio. Fortunately, the TTL output has typical rise and decay times of 10 nanoseconds.

The waveform may be selected from sine, triangle, or square, and these may be varied to pulse or sawtooth outputs by the variable duty-factor control.

Distortion is typically below 1% on the 0.2 Hz to 200 kHz ranges; outside these, it can rise to 4%. It was particularly noticeable at frequencies above 1.9 MHz, where a sine wave bears a distinct resemblance to a triangular one. Despite this, the waveform is still of reasonable purity and should prove acceptable for most purposes.

Symmetry is good over all ranges.

The output level is flat up to about 1.2 MHz when a slight decrease in amplitude occurs which continues up to the maximum output frequency. At 2 MHz the decrease amounts to 0.8 dB and at the maximum output frequency to 0.95 dB.

When the −26 dB attenuator is not in use, some distortion occurs at output levels below 2 V_{pp}. When the attenuator is in circuit, the output level can be reduced to 40 mV_{pp} before serious distortion sets in. Despite these few failings, the overall distortion of the instrument is well within specification and compares well with that of most other generators in this price range.

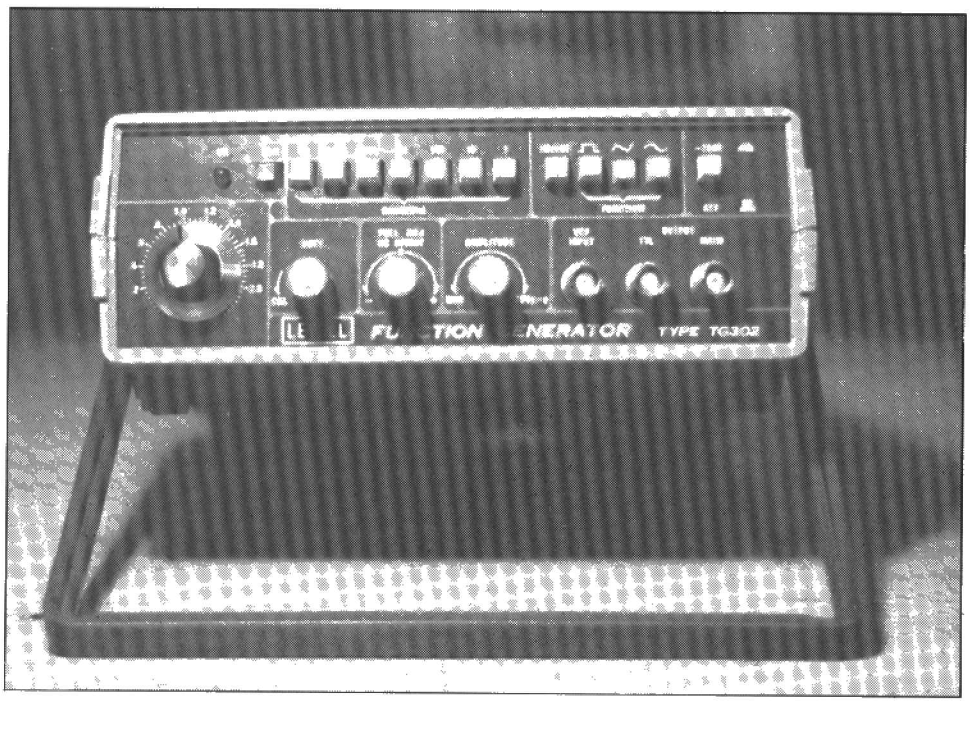
Frequency stability is good, although a noticeable change occurs when the output amplitude is varied. A VCF input facility is provided, which enables the frequency to be swept either up (positive input) or down (negative input) by up to three decades (1,000:1). The start frequency is set by the multiplying coefficient, and the off-set by the input voltage: a potential difference of 10 V gives a 1,000:1 change in frequency.

Although only 18 pages long, the manual contains some very helpful service, initial set-up, and calibration information, in addition to a good circuit description and circuit diagram.

Internal construction is based on a single, silk-screened, glass-fibre PCB on which all the main components, such as the mains transformer, power supply, generator circuitry, are mounted. The construction is very neat with a minimum of connecting wires to the various potentiometers and output sockets. The board is easily removed, and this should make servicing straightforward.

There is a fair amount of heat generated around the mains transformer and regulator and, although this should not present a serious problem, it should be taken into consideration if the instrument is to be used for long periods at a time.

The external casing is constructed from a two-part man-made-fibre enclosure and should allow the instrument to be used in most conditions. The swivel stand appears to be rather less robust.



Conclusion

At £136, excl. VAT, the TG302 should prove to be good value to most users who require a reasonably-well-built generator with good performance. The instrument has some useful facilities, such as the variable symmetry and the invert function. The manual is well prepared and should be particularly helpful for the servicing engineer.

On the negative side, distortion at higher frequencies and low output levels is rather higher than would normally be expected.

The TG302 was supplied by Levell Electronics Ltd, Moxon Street, BARNET EN5 5SD, telephone 01-440 8686.

Table 9

OPERATING RANGE

Frequency range: <0.02 Hz to 2 MHz in 7 calibrated ranges: fine adjustment by vernier, calibrated from 0.2× to 2× main setting.
Frequency accuracy: ±5% of full scale.
External sweep range: variable over >1000:1 ratio (>100:1 lowest range) by 10 V_{pp}.
Input impedance: 12 kΩ
Maximum slew rate: 0.2 V/μs.

OPERATING MODES

Sine wave: distortion <1% from 0.2 Hz to 200 kHz; harmonics <4% above 200 kHz; amplitude flatness <±0.1 dB up to 200 kHz, <±0.5 dB from 200 kHz to 2 MHz.
Square wave: mark:space ratio 1:1 ±1.5%.
DC range: ±5 V into 50Ω.
DC off-set: variable ±5 V into 50Ω.
Variable symmetry: ratio 1:15 or 15:1 by invert switch.

OUTPUTS

50Ω: <20 mV_{pp} to 20 V_{pp} from 50Ω ±1% source; <10 mV_{pp} to 10 V across 50 Ω load; switched attenuator reduces signal and d.c. off-set by 26 dB; output protected against short-circuits.
TTL: capable of driving up to 20 standard TTL loads.

GENERAL
Mains voltage: 110-120-220-240 VAC 50/60 Hz internally adjustable.
Power consumption: 10 VA.
Dimensions: 235×280×85 mm (W×D×H).
Weight: 2.0 kg.
Accessories supplied: mains lead (IEC terminated); manual; 2 test leads.
Warranty: 1 year.

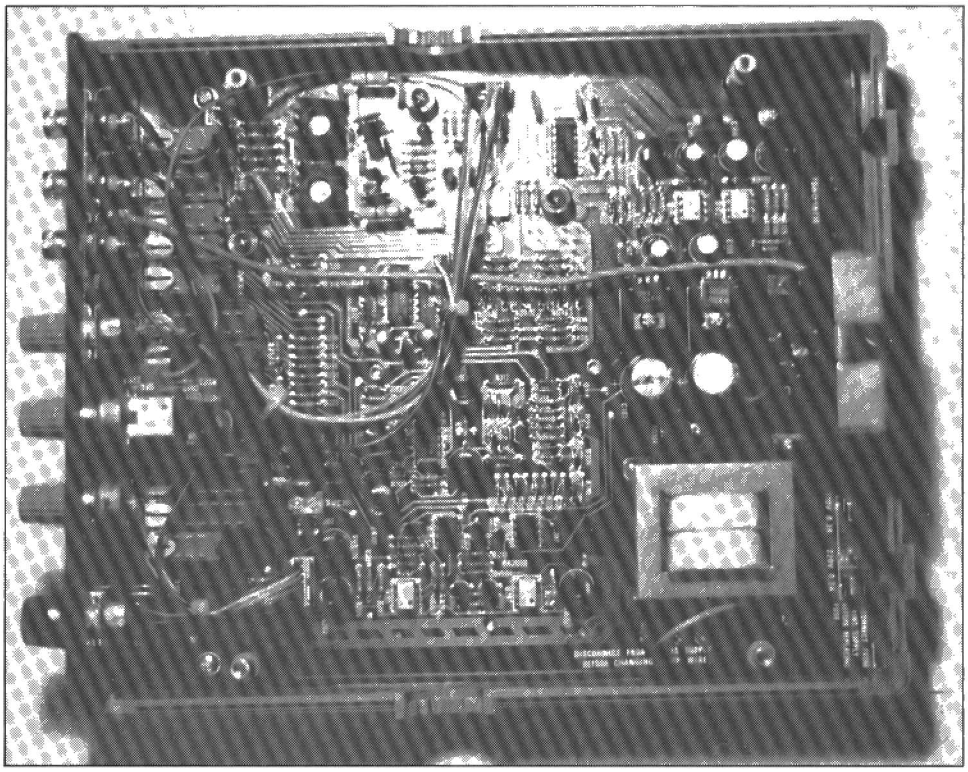


Table 10

	Unsatis-factory	Satis-factory	Good	Very good	Excellent
Dial accuracy				x	
Dial resolution			x		
External sweep range				x	
Distortion		x			
Frequency range				x	
Output level range					x
Internal construction				x	
External construction				x	
Overall specification				x	
Ease of use				x	
Manual				x	

Other signal generators available in the Levell range

TG152 – Low-distortion sine-square wave generator; frequency range 3 Hz to 300 kHz in five ranges; maximum output amplitude 2.5 V r.m.s. (7 V_{pp} on sine wave); synchronization output; battery or adaptor operation; high-accuracy setting of output possible; TG152D (no output level meter): £110 excl. VAT; TG152DM (output level meter fitted): £139 excl. VAT.

TG200 – Low distortion sine-square wave generator; frequency range 1 Hz to 1 MHz in 12 ranges; maximum output amplitude 7 V r.m.s. (20 V_{pp} on sine wave) can be reduced to 200 μV in seven switched steps; synchronization output; battery or adaptor operation; £149 to £185, excl. VAT, depending on options.

TG303 – as TG302 but includes 10 MHz counter and CMOS output; £236 excl. VAT.

Black Star Jupiter 2000

The Jupiter 2000 complements the already well-established Jupiter 500. Other Black Star products include a comprehensive range of frequency counters, digital multimeter, and a colour TV pattern generator.
The overall specification of the Jupiter 2000 is typical of a generator in this price range with a 2 MHz maximum output frequency, choice of 3 output waveforms, and variable symmetry. Not so typical are the 50-ohm and 600-ohm outputs and a switchable 0 dB, −20 dB, and −40 dB attenuator.
The front panel layout is clear and unambiguous with all the controls grouped neatly in sections for frequency, function, attenuation, and non-switched controls. The standard complement of seven overlapping frequency ranges is provided and these cover calibrated frequencies from 0.2 Hz to 2 MHz with a 10:1 vernier. Overall calibration accuracy of the vernier is reasonable and remains within the ±5% error limit over all frequency ranges.

Mains voltage is selected internally. The instrument is connected to the mains by an IEC connector. Unusually, the on-off switch is at the rear of the generator, which, although helping to maintain the clear front panel layout does not enhance the Jupiter 2000's ease of operation if it is stacked or of necessity placed with the rear panel inaccessible.
The instrument is fitted with a one-position stand, which enables the instrument to be used at optimum operating angle.
Unusually, the output frequency is selected on a linear scale. The multiplying coefficient is variable over a calibrated range of 0.2 to 2 times the range setting. One of the Jupiter's main assets must be its frequency stability over all multiplying coefficients. In contrast to many other generators in this price range, the output frequency is maintained at constant level down to the lowest frequency of 0.09 Hz. While this minimum output frequency is unlikely to be a serious disadvantage, it is higher than the 0.02 Hz offered on some rival instruments.
Symmetry is variable over a range of 8:1 on sine, triangle, and square wave outputs. Overall, this function performs very well, despite its being only usable to a maximum frequency of 200 kHz (which is due to the output frequency being divided by 10 when the symmetry function is actuated). Since the variable symmetry is probably most significant at lower frequencies, this may not be too important, unless the facility is specifically required up to the maximum output frequency.
Linearity is good on all ranges and multiplying coefficients, well within the quoted ±1% up to about 1 MHz (square wave).
The rise time on the TTL output is typically 15 nanoseconds and the decay time 10 nanoseconds. Rise time on the main 50-ohm and 600-ohm outputs is 70 ns.
The overall frequency accuracy is good with the vernier allowing precise settings to a high resolution over the whole range. The slightly unusual step of using a linear control here has both advantages and disadvantages in the accuracy of setting at high and low multiplying coefficients respectively.
Waveform is selected from sine, triangle, and square wave. On the whole, distortion on all ranges is low, typically below 0.8% to 200 kHz (sine wave). There is, however, some very slight distortion on each of the positive and negative peaks (sine wave) on some ranges, although in the main this is insignificant and does not degrade the Jupiter 2000's good overall performance.
The output level is reasonably flat with a typical variation of ±0.18 dB to 200 kHz and ±0.9 dB to the maximum output frequency of 2 MHz.
One of the main features is the instrument's versatile output attenuator. This can select three levels of attenuation, which makes low output levels easily obtainable, while good maximum output levels of 20 V_{pp}, 2 V_{pp}, or 0.2 V_{pp} are maintained.
Noise is kept to a low level over all output ranges, so that a realistic minimum output level of 2 mV can be achieved with the attenuator in circuit. No increase in distortion is apparent at relatively low output levels.
The effect of the d.c. off-set control, which usually gives an off-set of up to ±10 V (0 dB), is reduced proportionally in the −20 dB and −40 dB positions.
There are three outputs available on the instrument: 50 ohms, 600 ohms, and TTL. The provision of two main outputs increases the ease of use, which is often reduced in instruments with only the usual 50-ohm output.

Sweep

Linear sweep ranges of over 1000:1 can be obtained on the 1 kHz, 10 kHz, 100 kHz, and 1 MHz ranges, while on the 100 Hz range the sweep range is restricted to typically 130:1, and on the 1 Hz and 10 Hz ranges to about 35:1.
Non-linear sweeps can be carried out with ratios of up to 2,000:1, which increases the maximum output frequency to 4 MHz, although, obviously, level flatness and distortion are far from good at this frequency.
Input sensitivity is slightly low at 17.1 V_{pp} or 19.1 V_{pp} for 10:1 and 2,000:1 sweeps respectively, but this is compensated to some extent by the instrument's above-average maximum input slew rate of 0.1 V/μs.

Manual

The 7-page manual contains brief and to-the-point operational information, and covers sweep operation in some detail. However, a circuit diagram or description, and calibration or service information are not included. A service manual is available separately.

Table 11

OPERATING RANGE

Frequency range: <0.2 Hz to 2 MHz in 7 calibrated ranges: fine adjustment by vernier, calibrated from 0.2× to 2× main setting.
Frequency accuracy: ±5% of full scale on all ranges.
External sweep range: variable over >1000:1 ratio (>100:1 lowest range) by 19 V_{pp}.
Input impedance: 9 kΩ
Maximum input: ±50 V_{pp}.

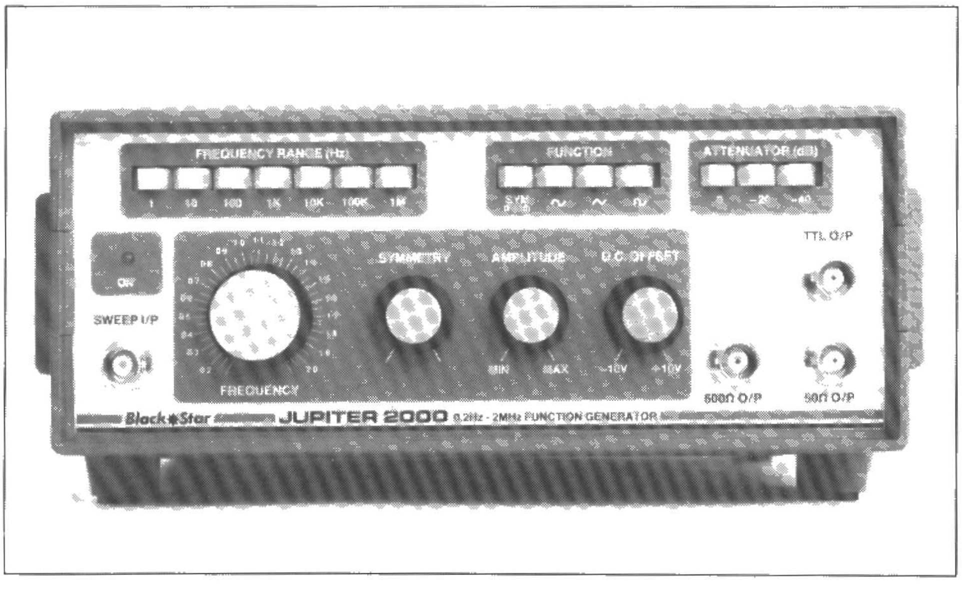
OPERATING MODES

Sine wave: distortion <0.5% from 0,2 Hz to 200 kHz; <1% from 200 kHz to 2 MHz; harmonics >30 dB below fundamental.
Square wave: mark:space ratio 1:1 ±1% to 200 kHz.
DC range: ±10 V into 50Ω.
DC off-set: variable ±5 V into 50Ω.

OUTPUTS

50Ω: <0 to 20 V_{pp} from 50Ω ±1% source; 0 to 10 V across 50 Ω load; 0 to 10 V_{pp} across 600-ohm load; switched attenuator reduces signal and d.c. off-set by 0 dB, 20 dB or 40 dB; output protected against short-circuits.
TTL: capable of driving up to 20 standard TTL loads.

GENERAL
Mains voltage: 110-120-220-240 VAC 50/60 Hz internally adjustable.
Power consumption: 24 VA.
Dimensions: 219×240×98 mm (W×D×H).
Weight: 1.6 kg.
Accessories supplied: mains lead (IEC terminated); manual.
Warranty: 1 year.



Construction

The instrument is based on one double-sided PCB, which contains all the main components, although larger ones are mounted on the chassis. Interconnection between the board and chassis is in some instances by ribbon cable.

Generally, the construction is of a high standard, and the quality of finish is first class throughout.

Accessibility is good in all respects and this should make servicing fairly easy.

External construction is based on a two-part man-made-fibre enclosure with aluminium end plates. The whole should prove sufficiently rugged for most applications.

Table 12

	Unsatis-factory	Satis-factory	Good	Very good	Excellent
Dial accuracy				x	
Dial resolution				x	
External sweep range				x	
Distortion			x		
Frequency range				x	
Output level range					x
Internal construction					x
External construction				x	
Overall specification					x
Ease of use				x	
Manual			x		

Conclusion

At £149 (excl. VAT), the Jupiter 2000 is good value for money, offering a high standard of construction and a good standard of performance and facilities. Compared with other instruments in this price range, the Jupiter 2000's choice of 50-ohm or 600-ohm outputs and three attenuation ranges make it worth special consideration.

The two main quirks of the generator are its linear frequency control (which, however, has its good and bad points), and the slight distortion on the peaks of the sine wave output. The latter should, however, not be of much consequence to most users who require a general-purpose function generator.

The instrument offers a wide range of facilities and these more than offset the slight failings just mentioned.

The Jupiter 2000 was supplied by Black Star Ltd, 4 Harding Way, St. Ives, HUNTINGDON PE17 4WR

Other signal sources available from Black Star

Jupiter 500 – 0.1 Hz to 500 kHz function generator; external AM and sweep facilities; distortion typically 0.5% to 100 kHz; 30 V_{PP} output amplitude; −20 dB attenuator; £110.00 excl. VAT.

Orion PAL colour bar/video pattern generator – price £199.00 excl. VAT.

REAL-TIME CLOCK PATCH FOR DCF77 ON THE COMMODORE C64

An experimental interrupt routine is discussed that enables the Commodore C64 home computer to display time and date information received from time standard service DCF77.

The publication in this magazine of a receiver, intelligent time standard, and locked frequency reference based on Federal Germany's time signal transmitter DCF77 (References 1. and 2.) has been a great success. Usable reception for synchronizing the microprocessor-driven clock in the intelligent time standard has been reported by numerous constructors in the UK, Ireland and Europe. Long-distance reception with synchronization taking place up to 10 times a day was reported from as far as Saudi-Arabia and central Sweden. Clearly, the Deutsche Bundespost has remained on the safe side when claiming that the transmitter's normal range is about 800 kilometres.

The DCF77 receiver and locked frequency standard was designed to operate in conjunction with the 8052AH-BASIC driven intelligent time standard. Since this is essentially a small microcomputer system that translates received time pulses in a time and date indication, it would appear logical to investigate how other computers can be programmed to accomplish this, keeping in mind that all the advanced features of the intelligent time standard will prove very difficult, if not impossible, to implement on a simple home computer such as the C64. From the onset it should be made clear that the program discussed here is experimental, and, therefore, aimed at C64 users thoroughly familiar with programming this home computer.

The real-time clock program is in fact an interrupt request routine. This means that it will work only in conjunction with programs that do not use interrupts. Most games for the C64 do, and can not, therefore, use the present clock program.

The RTC routine resides in RAM between CC00H and CF00H. One peculiarity of the C64 is that interrupts are sometimes disabled, for instance, during disk operations. The result is that the clock is not updated temporarily, so that the time indication on screen does not correspond to the time transmitted by DCF77 (consult Ref. 2. for the structure of the time signals). Synchronization will be restored, however, within a minute after the enabling of interrupts.

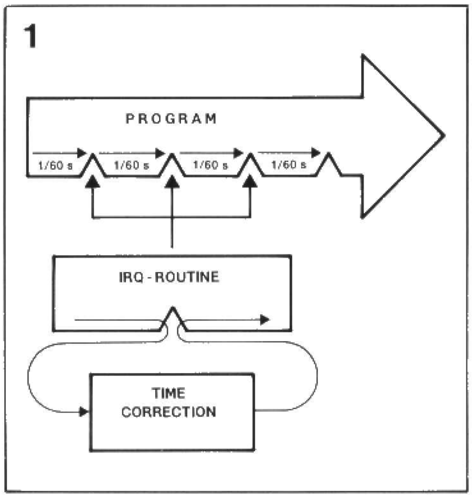


Fig. 1. The main program running on the computer is interrupted at a rate of 60 Hz for reading and processing the time pulses from DCF77. The IRQ start vector is redirected by an initialization routine (SYS 52224).

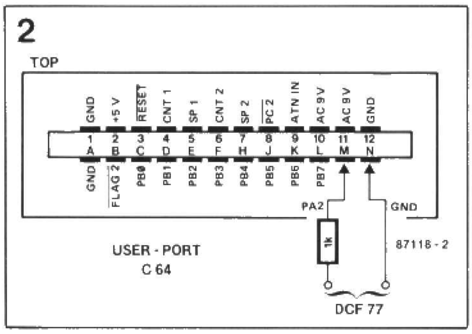


Fig. 2. Connection of the DCF77 receiver to the User Port at the rear of the C64.

As already stated, the program is of interest to programmers who know their way on the C64. As such, it is a guide to further experiments, and definitely not an attempt to imitate the many powerful features of the control program in the intelligent time standard.

Interrupt handling

The drawing in Fig. 1 illustrates the basic operation of the time correction procedure on the C64. The main program is interrupted 60 times per second to analyse the structure of the received time pulses. C64 programmers are no doubt familiar with the two CIA6526s which have a built-in real time clock (RTC). The RTC in CIA2 is used here to keep the clock running when the signal from DCF77 is temporarily lost.

Figure 2 shows that the time signals provided by the DCF77 receiver and locked frequency standard are fed to the C64 via User Port line PA2. This is programmed to function as an input line each time the processor enters the interrupt routine. Some programs use the PA2 line as a serial output, so that the CIA must be protected by a current limiting resistor as shown.

Operation

The RTC program uses only the minute, hour, day, month and year information transmitted by DCF77. The absence of a signal to mark the last second in each minute enables the program to recognize the start of the next minute. Next, the processor awaits the 21st second before updating the binary coded hour and date information on the basis of the received bit combinations. Logic bit levels are deduced from the pulse-width of the time signal, 100 ms corresponding to logic 0, and 200 ms to logic 1. The results of the bit-processing routines are not sent to the screen before the absence of the code for the 59th second is detected. At the same time, the RTC in CIA2 is synchronized. The sign @ appears on the screen in the location of hour and date to indicate that the RTC in CIA2 is not yet synchronized.

Error codes 1, 2, 3 or 4 may appear on the screen in inverse video. Their meaning is as follows:

1 = data pulse is too long;
2 = the difference between two successive minutes is other than 1;
3 = data is not coherent;
4 = synchronization lost (59th second missed or prematurely recognized).

The program has been simplified to keep it as short as possible. It has neither parity checking nor verification of received data at the beginning of the hour.

References:

1. DCF77 receiver and locked frequency standard. *Elektor Electronics* January 1988, p. 24 ff.

2. Intelligent time standard. *Elektor Electronics* February 1988, p. 22 ff.

```
 5 REM DCF77 ON COMMODORE c64
 6 REM RTC PATCH FOR ELEKTOR ELECTRONICS INTELLIGENT TIME STANDARD
10 N=0:FORI=0TO046:FORK=0TO16:READX:N=N+X:POKE52224+I*17+K,X:NEXTK:READX
20 IF X<>N THEN PRINT"ENTRY ERROR IN LINE ";I+40:END
30 N=0:NEXTI:SYS52224
40 DATA76,3,204,120,173,15,221,41,127,141,15,221,173,14,221,9,128,1902
41 DATA141,14,221,169,0,141,11,221,141,10,221,141,9,221,141,8,221,2031
42 DATA162,32,157,23,207,202,16,250,169,59,141,20,3,169,204,141,21,1976
43 DATA3,169,32,141,47,207,88,96,169,59,141,2,221,173,51,207,208,2014
44 DATA3,76,209,206,173,23,207,201,21,48,9,201,58,48,66,208,3,1760
45 DATA76,10,205,173,48,207,240,28,173,0,221,41,4,208,31,173,52,1890
46 DATA207,201,50,48,24,169,0,238,23,207,141,48,207,141,52,207,76,2039
47 DATA82,205,173,0,221,41,4,240,3,141,48,207,238,52,207,173,52,2087
48 DATA207,201,70,48,233,169,0,141,50,207,76,222,205,173,48,207,208,2465
49 DATA195,173,0,221,41,4,240,226,141,48,207,173,52,207,201,8,48,2185
50 DATA15,201,14,48,8,169,177,141,47,207,76,141,204,56,176,1,24,1705
51 DATA110,49,207,173,23,207,201,28,208,11,173,49,207,41,127,141,27,1982
52 DATA207,76,49,234,201,36,208,11,173,49,207,41,63,141,31,207,76,2010
53 DATA49,234,201,43,208,11,173,49,207,41,63,141,35,207,76,49,234,2021
54 DATA201,52,208,11,173,49,207,41,31,141,39,207,76,49,234,201,57,1977
55 DATA208,6,173,49,207,141,43,207,76,49,234,238,52,207,173,48,207,2318
56 DATA208,14,173,0,221,41,4,240,54,169,7,141,48,207,208,47,173,1955
57 DATA0,221,41,4,208,25,169,0,141,51,207,141,48,207,141,23,207,1834
58 DATA141,52,207,141,54,207,169,180,141,47,207,76,141,204,173,52,207,2399
59 DATA201,64,208,8,238,23,207,169,0,141,52,207,76,222,205,173,23,2217
60 DATA207,201,60,240,3,76,222,205,169,0,141,23,207,173,50,207,240,2424
61 DATA7,173,27,207,240,9,208,36,173,27,207,201,89,240,229,169,1,2243
62 DATA141,50,207,32,127,205,76,222,205,160,5,162,0,189,27,207,232,2247
63 DATA157,27,207,232,232,232,136,208,243,96,248,24,237,28,207,216,170,2900
64 DATA240,8,169,178,141,47,207,76,141,204,160,4,162,0,189,31,207,2164
65 DATA232,221,31,207,240,8,169,179,141,47,207,76,141,204,232,232,232,2799
66 DATA136,208,233,32,127,205,169,2,141,50,207,169,32,141,47,207,173,2279
67 DATA32,207,141,11,221,173,28,207,141,10,221,169,0,141,9,221,141,2073
68 DATA8,221,173,50,207,201,2,240,71,173,8,221,240,3,76,112,206,2212
69 DATA173,11,221,141,53,207,41,15,9,48,141,33,207,32,3,207,141,1683
70 DATA34,207,173,10,221,141,53,207,41,15,9,48,141,29,207,32,3,1571
71 DATA207,141,30,207,173,9,221,141,53,207,41,15,9,48,141,25,207,1875
72 DATA32,3,207,141,26,207,173,8,221,76,112,206,173,23,207,240,25,2080
73 DATA162,0,201,10,144,5,233,10,232,208,247,9,48,141,25,207,138,2020
74 DATA9,48,141,26,207,76,112,206,169,48,141,25,207,141,26,207,160,1949
75 DATA5,162,0,189,28,207,141,53,207,41,15,9,48,157,29,207,32,1530
76 DATA3,207,157,30,207,232,232,232,232,136,208,230,173,25,207,141,39,2691
77 DATA4,173,26,207,141,38,4,169,58,141,37,4,141,34,4,173,29,1383
78 DATA207,141,36,4,173,30,207,141,35,4,173,33,207,141,33,4,173,1742
79 DATA34,207,141,32,4,173,45,207,141,79,4,173,46,207,141,78,4,1716
80 DATA169,45,141,77,4,141,74,4,173,41,207,141,76,4,173,42,207,1719
81 DATA141,75,4,173,37,207,141,73,4,173,38,207,141,72,4,173,47,1710
82 DATA207,141,30,4,76,49,234,173,0,221,41,4,208,32,169,85,141,1815
83 DATA54,207,173,52,207,240,11,201,70,48,7,201,130,16,3,141,51,1812
84 DATA207,169,0,141,52,207,141,23,207,76,222,205,173,54,207,240,3,2327
85 DATA238,52,207,76,222,205,110,53,207,110,53,207,110,53,207,110,53,2273
86 DATA207,173,53,207,41,15,9,48,96,7,0,57,48,25,24,48,50,1108

READY.
```

Listing of the real-time clock routine that enables the Commodore C64 to display time and date information received from DCF77.

EVENTS

IEE Meetings

1 – 3 July **History of electrical engineering**

10 – 15 July **Local telecommunications** - Fourth vacation school at the University of Aston.

11 – 15 July **Software engineering 88** - Exhibition and Conference at the University of Liverpool.

18 – 20 July **Third International Conference on Image Processing and its Applications** at the University of Warwick.

24 – 29 July **Optical fibre communications** - Fourth vacation school at the University College of Bangor.

31 July – 4 Aug **Satellite communications** - Fourth vacation school at the University of Surrey.

4 – 9 Sept **Switching and signalling in telecommunication networks** - Ninth vacation school at the University of Aston.

4 – 10 Sept **Measurement technology (DC to VHF)** - Vacation school at the University of Aston.

11 – 15 Sept **Optical communication (ECOC)** - Fourteenth European conference and exhibition in Brighton.

11 – 16 Sept **Telecommunication network design and performance** - Vacation school at the University of Strathclyde.

23 – 27 Sept **International Broadcasting Convention (IBC)** - Conference and exhibition in Brighton.

Further details on the above events may be obtained from the Secretary, IEE, Savoy Place, LONDON WC2R 0BL, telephone 01-240 1871.

CONPAR 88 will be held on 10 – 16 Sept in Manchester. Further details from the British Computer Society, 13 Mansfield Street, LONDON W1M 0BP, telephone 01-637 0471.

The sixth international conference on **Electromagnetic compatibility** will be held at the University of York on 11 – 15 Sept. Further details from the Institution of Electronic and Radio Engineers, Savoy Hill House, Savoy Hill, LONDON WC2R 0JD, telephone 01-240 1871, Ext. 246.

An **Artificial intelligence in real-time control** workshop will be held at the University College of Swansea on 21 – 24 Sept. Further details from the Institute of Measurement and Control, 87 Gower Street, LONDON WC1E 6AA, telephone 01-387 4949.

The **Electronics India Exhibition** will be held in New Delhi on 6 – 11 Sept. Further information from GAMBICA, 8 Leicester Street, LONDON WC2H 7BN, telephone 01-437 0678.

The **World Administrative Radio Conference** on the use of the geostationary-satellite orbit and on the planning of space services utilizing it will be held in Geneva from 29 August to 5 October.

Leetronex, the Leeds Electronics Exhibition, will be held at the University of Leeds (Electrical and Electronic Engineering Department: telephone (0532) 420339) on July 5 – 7.

A **Mobile Radio Communications Exhibition** will be held at Sandown Park, Esher, on 13 – 15 Sept. Further details from Framework, telephone 01-778 5656.

The **Electronics Industry Exhibition** will be held in Hong Kong on 15 – 18 Sept. Further details from ADG Exhibitions, telephone (0243) 29406.

The **Canadian High Technology Week** will be held in Toronto on 27 – 29 Sept. Further details from Overseas Trade Agencies, telephone 01-486 1951.

The **Semiconductor International Exhibition** will be held at the NEC, Birmingham, on 27-29 Sept. Further information from Cahners Exhibitions, Chatsworth House, 59 London Road, TWICKENHAM TW1 3SZ, telephone 01-891 5051.

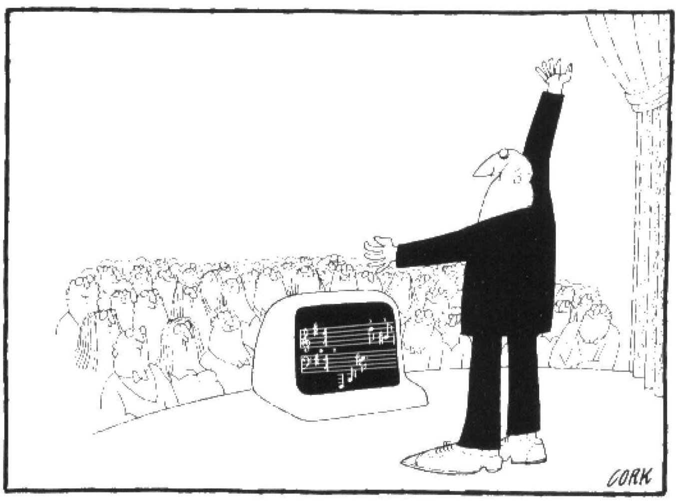
Test and Transducer International, the instrumentation Exhibition and Conference, will be held at the NEC, Birmingham, on 27 – 30 Sept. Further information from Trident International Exhibitions Ltd, 21 Plymouth Road, TAVISTOCK PL19 8AU, telephone (0822) 4671.

An ERA Technology seminar on **Defence Standard 59 – 41**, which lays down the electromagnetic compatibility (EMC) requirements for all equipment and systems used by the MoD, will be held at the Cavendish Conference Centre, London, on 7 July. Further information from ERA Technology Ltd, Cleeve Road, LEATHERHEAD KT22 7SA, telephone (0372) 374151.

The **Light and Sound Show** (sound, lighting, video, lasers, equipment and accessories for leisure industries) will be held at Olympia 2, London, on 11 – 14 Sept. Further information from the Professional Lighting and Sound Association, 1 West Ruislip Station, RUISLIP WA4 7DW, telephone (08956) 34515.

ERA Technology is to organize a **Conference on coil winding** which will run in tandem with the **Coil Winding International Exhibition** being staged by Evan Steadman (Services) Ltd on 6 – 8 Sept at the Wembley Conference Centre, London. Further information from ERA Technology Ltd, Cleeve Road, LEATHERHEAD KT22 7SA, telephone (0372) 374151, or Evan Steadman (Services) Ltd, The Hub, Emson Close, SAFFRON WALDEN CB10 1HL, telephone (0779) 26699.

This year's **EPoS/EFTPoS (Electronics at the point of-sales/electronics funds transfer at the point-of-sale)** congress will be held at the newly refurbished Alexandra Palace, London, on 13 – 16 Sept. Further information from RDMP Ltd, 61 Ship Street, BRIGHTON BN1 1AE.

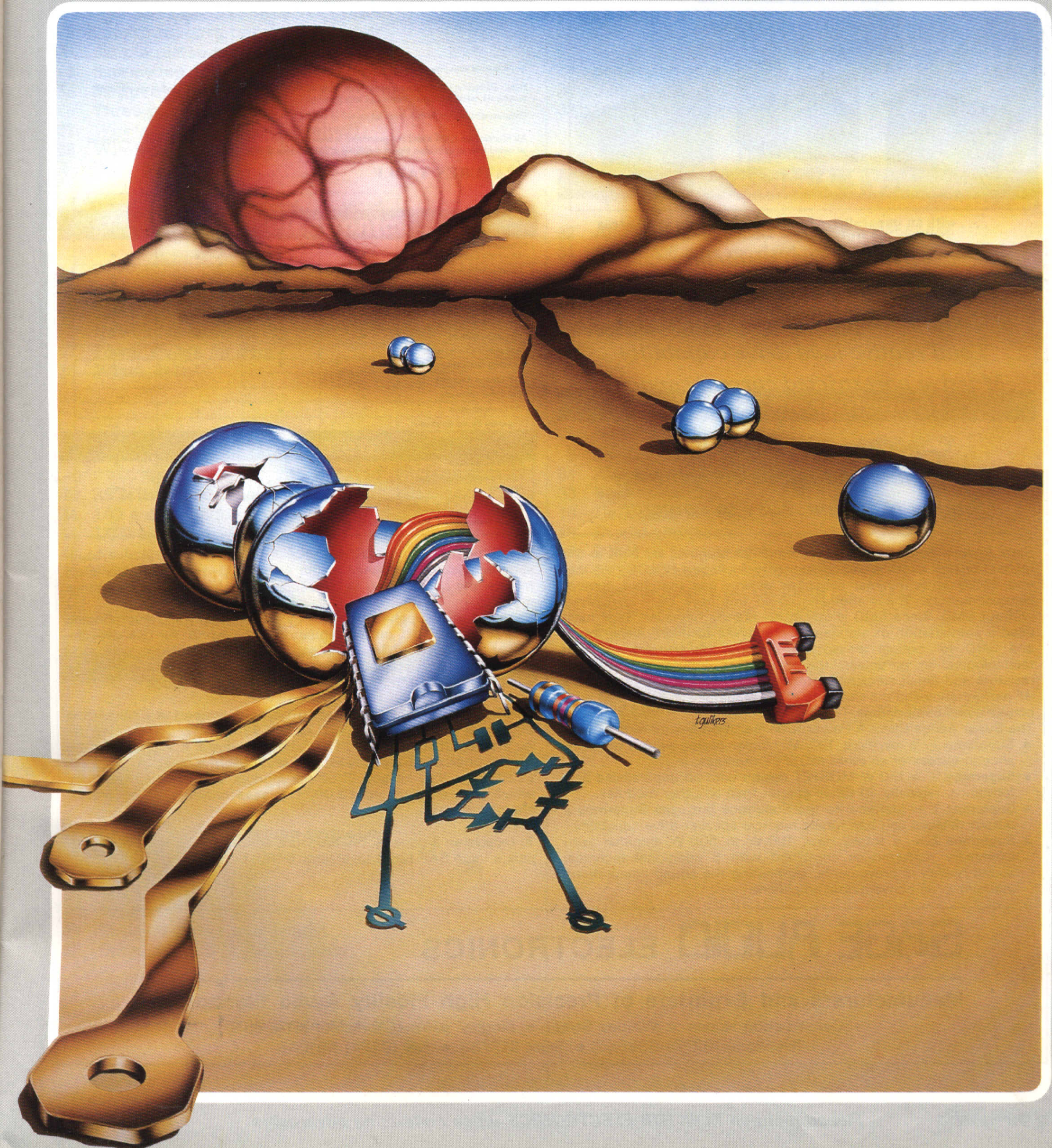


THE PRACTICAL MAGAZINE WITH THE PROFESSIONAL APPROACH

July/August 1988

Elektor Electronics

Supplement: a miscellany of construction projects



Elektor Electronics

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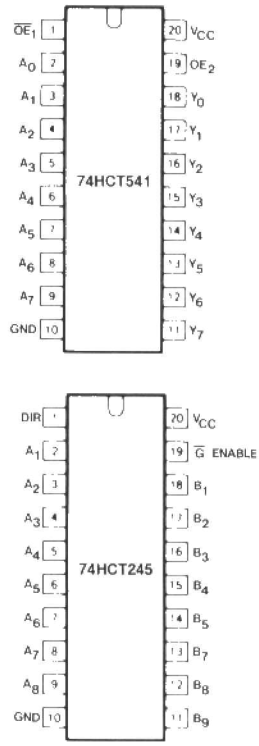
001 PROTOTYPING BOARD FOR COMPUTER EXTENSIONS

This printed circuit board is ideal for building and testing experimental extension circuits for a wide range of computers. The double-sided, but not through-plated, board has contact fingers that enable it to be accepted in commonly used slot connectors for extension circuits in many types of computer, including those in the MSX and IBM PC series. In addition, the board holds 3 general-purpose buffer chips which can be wired to requirement to ensure correct and safe interfacing between the computer and the extension circuit being developed. Supply tracks are provided in the buffer and prototyping area on the board for ease of wiring. When required, a number of contact fingers can be cut off to suit a particular slot width, or to prevent the board being fitted the wrong way around in the slot. Also, the contact fingers are relatively long so that a section of this PCB area can be cut off for use as an adaptor together with a purchased slot connector. It is also possible to fit a slot connector at right angles at either side of the PCB as shown by the printed markers. The pin connections of the Type 74HCT245 octal transceiver, and the Type 74HCT541 octal three-state line buffer are given here for reference. These chips are suggested for use as databus and addressbus buffers respectively, because they have inputs and outputs arranged at opposite sides of the 20-way DIL enclosure. The user is, however, left completely free to choose his own bus buffers in accordance with the interfacing requirements. Remember to ground unused inputs on HCT chips!

Note: the printed circuit board for this project is available ready-made through the Readers Services as order no. 884013 (see Readers Services page).



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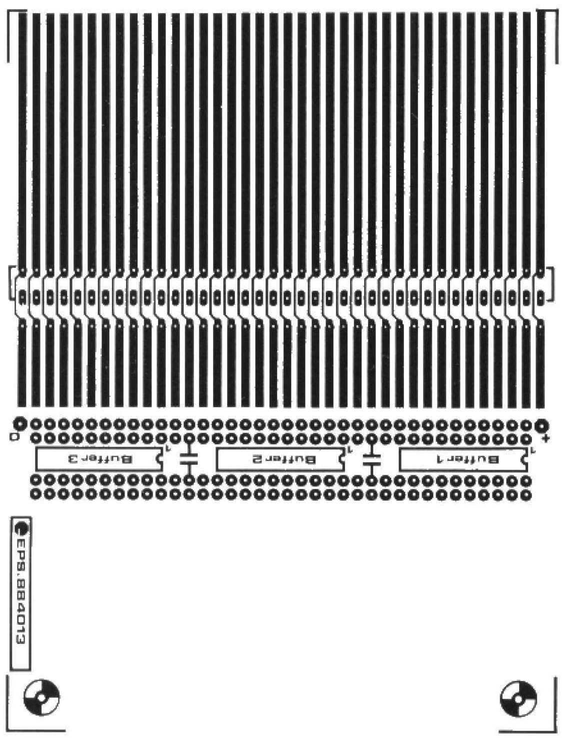
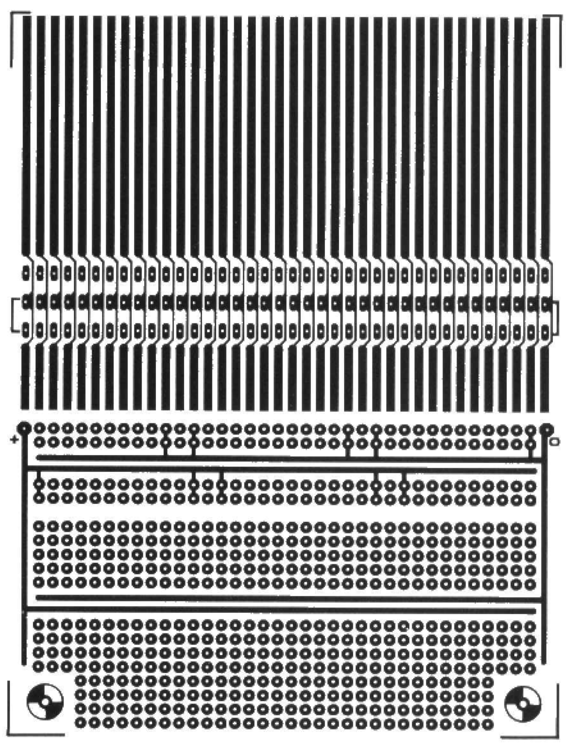
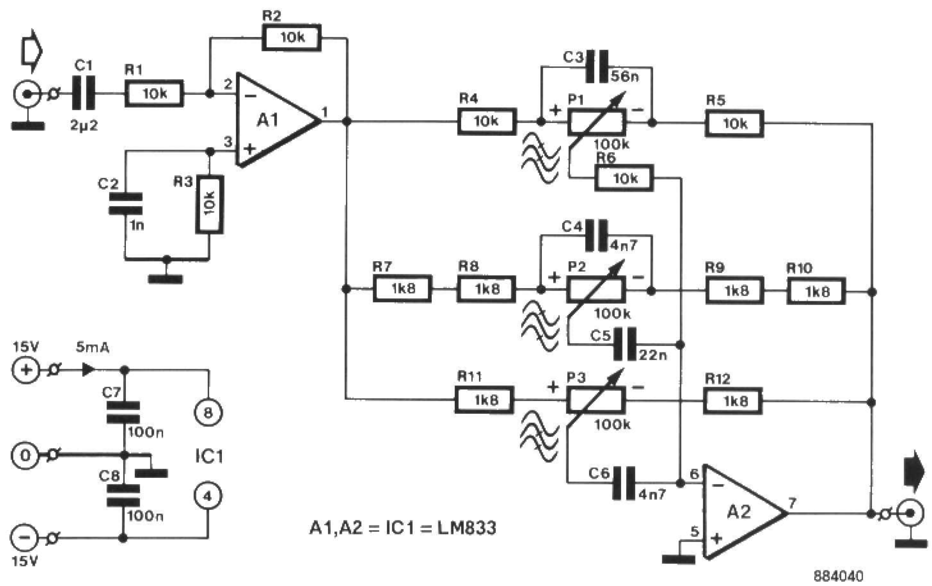




002 THREE-WAY TONE CONTROL

Although tone control is not desirable in good-quality audio equipment, there are still instances, such as when playing well-used records, when it is. Such an add-on tone control should enable the frequency response to be altered to taste, have no detrimental effect on the audio equipment, and be fairly compact. The circuit proposed here meets these criteria.
It is based on National Semiconductors's LM833. This dual operational amplifier has a very low noise factor (4.5 nV/$\sqrt{f}$ (Hz)), a high gain-bandwidth product (15 MHz), and a slew rate of 7 V/μs.
The tone control circuit consists of three ranges, so that a presence control at around 1 kHz is possible.
The opamp at the input, A_1, is connected as an inverting buffer. Its non-inverting input is connected to a 10 k resistor to equalize the direct currents at both inputs (with respect to the bias currents). This is necessary to keep the output of A_1 near enough at 0 V because of the d.c. coupling to A_2.
The second opamp has in its feedback loop a simple three-way tone control, whose cross-over points are determined by the value of the four capacitors.
If desired, a capacitor may be added to the output of A_2, because the d.c. output of this opamp varies somewhat with the setting of the potentiometers.

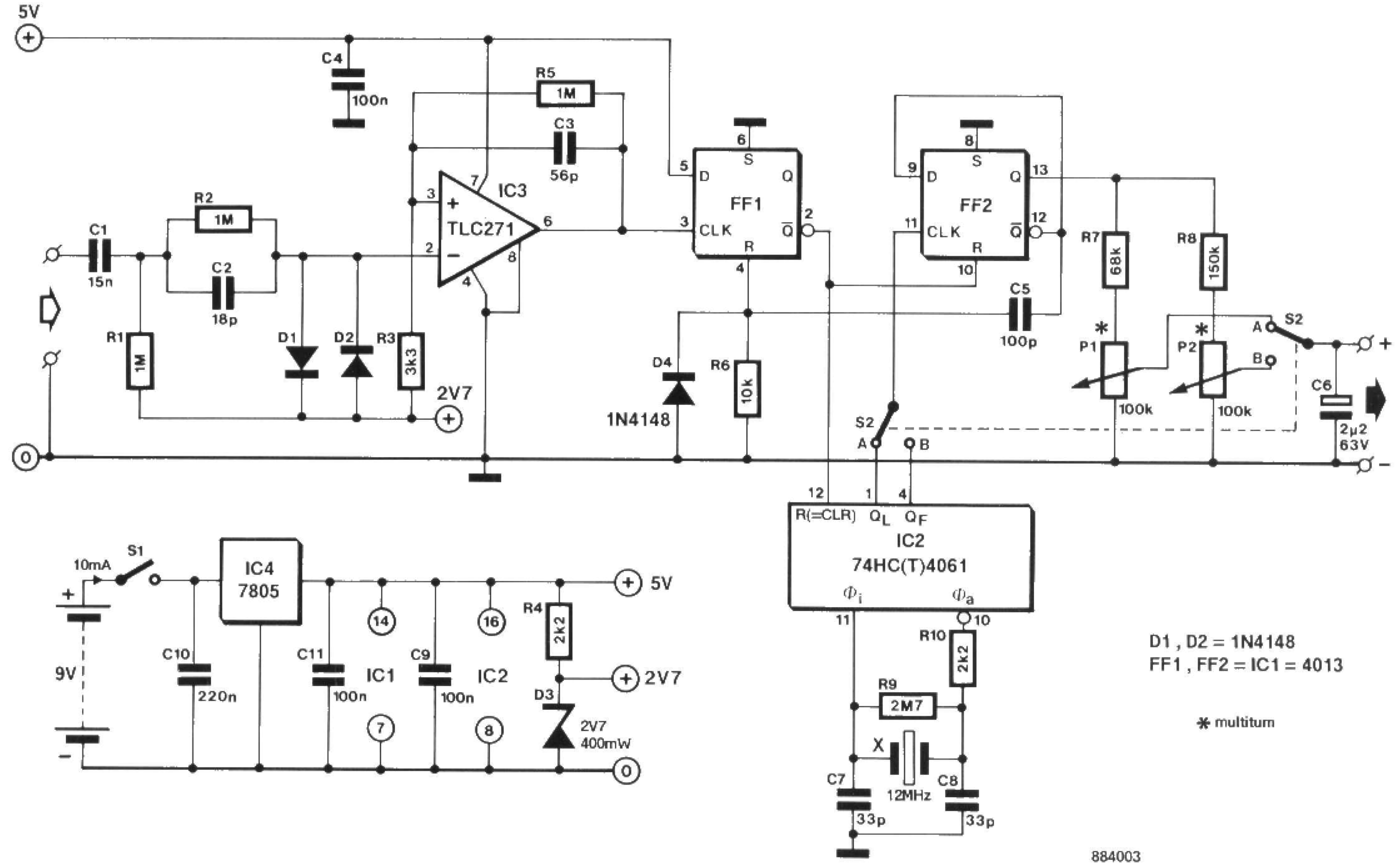


The cross-over points of the low-frequency and high-frequency controls lie at about 200 Hz and 2 kHz respectively. The presence control operates at around 1 kHz.
Maximum attenuation is about 16 dB.

With all potentiometers at the centre of their travel, the signal-to-noise ratio is better than 90 dB at a bandwidth of 1 MHz. The gain is 0 dB but can be altered by changing the value of R_2.

003 DMM AS FREQUENCY METER

By providing a high-input-resistance multimeter (preferably of the digital type) with a frequency-to-voltage converter, it can be used to measure frequency.
The range of the proposed device extends from 10 Hz to 1 kHz on range A and from 1 kHz to 100 kHz on range B.
The sensitivity for frequency measurements up to about 10 kHz is of the order of 35 mV_{pp}, and for measurements from 10 kHz to 100 kHz about 350 mV_{pp}.
The input signal is applied to Schmitt trigger IC_3 via limiters D_1 and D_2. Bistables FF_1 and FF_2, and IC_2 form a monostable. When the monostable is triggered, it generates a pulse whose width is accurately determined by a 12-MHz crystal.
The number of times the monostable is triggered per unit time depends on the input signal.
The pulse height depends on the supply of the monostable. The supply is provided by voltage regulator IC_4 and is about 5 V.
At the output of the monoflop, i.e., pin 13 of FF_2, there will thus be a train of pulses, whose width and height are constant, but whose number and, therefore, the average voltage is directly proportional to the input frequency.
The RC network at the output of FF_2 forms a low-pass filter, so that the average voltage of the pulses will appear across C_6.
Potentiometers P_1 and P_2 and resistors R_7 and R_8 form a potential divider which enables the frequency-to-voltage conversion factor to be adjusted.
The voltage across C_6 measured by the DMM is thus directly proportional to the frequency of the input signal.
In range A, a voltage of 10 mV corresponds to 10 Hz, and 1 V to 1 kHz. In range B, 10 mV corresponds to 1 kHz, and 1 V to 100 kHz.
For adjusting the meter, temporarily connect the junction of R_7 and R_8 to pin 12 instead of to pin 13 of FF_2. There should be no input signal. Set the DMM to the 20 V range, and connect it across C_6. Set S_2 to position A, and adjust P_1 until the meter reads 2.93 V. Then set the meter to the 2 V range, and S_2 to position B. Now adjust P_2 until the meter reads 1.875 V. Finally, reconnect the junction of R_7 and R_8 to pin 13 of FF_2.
The meter may be powered by a 9-V PP3 battery: the current consumption amounts to only 10 mA.



004 RTTY FILTER FOR 170 HZ SHIFT

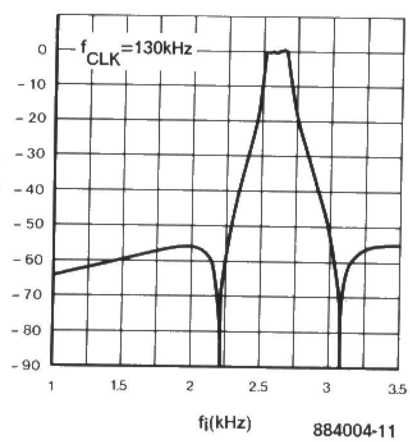
Anyone interested in the reception of radio teletype traffic will appreciate the programmable audio filter described here, which may be fitted at the input of the RTTY converter. It improves the signal/-to-noise ratio, particularly of signals in the crowded short-wave bands.

The circuit is based on programmable filter IC_3 – see Fig. 2. The special facet of this IC is that the resistors of the on-board RC filters are simulated by capacitors. This little-known technique is described in the January 1981, October 1982, and February 1983 issues of this magazine.

The value of the capacitors, and therefore the pass-band frequency of the filter, is determined by the frequency of the clock at pin 8 of IC_3. The clock frequency is made variable by passing a 10-MHz signal through a programmable divider, IC_1. The divisor may be set between 1 and 256 with the aid of switches S_{1a} – S_{1h}.

Monostable IC_2 converts the output pulses from the divider into a near-symmetrical signal, which is subsequently used as the clock for IC_3 (pin 8).

1

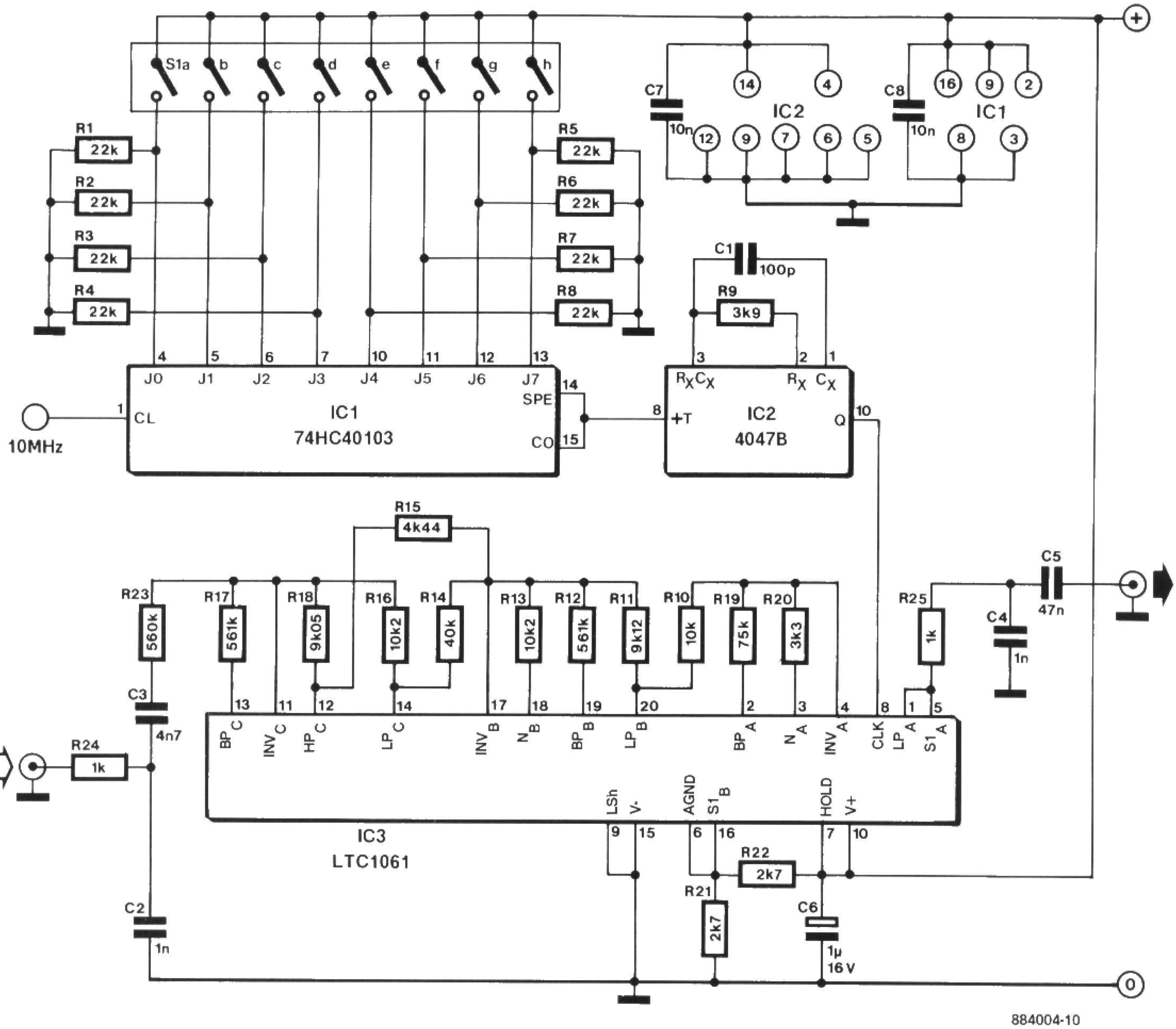


The filter characteristic is shaped by the resistors at the various pins of IC_3 to a very-narrow pass-band as is required for small RTTY shifts. The characteristic is shown in Fig. 1. The entire pass-band may be shifted with the aid of the switches.

In narrow-band RTTY (70 – 170 Hz shift), one filter suffices, since both the high and low AFSK frequencies can be passed by the filter. For broadband RTTY signals (425 – 850 Hz shift), it is probably better to use separate filters for the high and low frequencies.

The circuit draws a current not exceeding 20 mA.

2

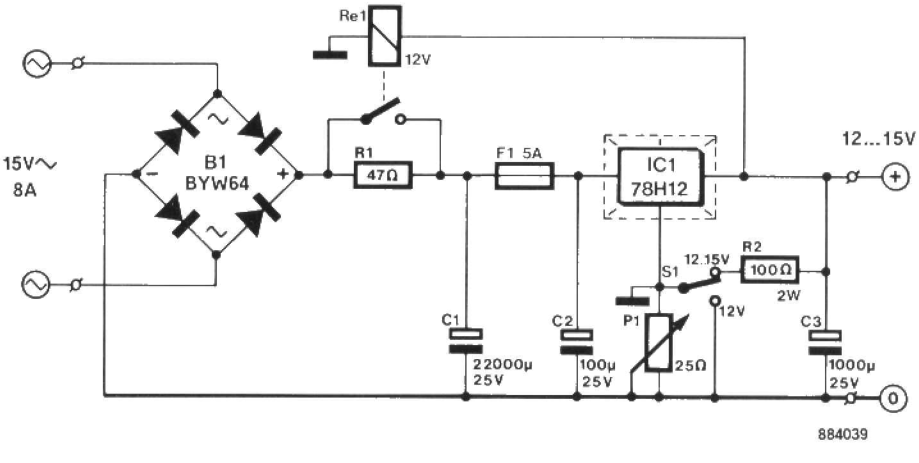


005 SECONDARY POWER-ON DELAY

The circuit described here enables short-circuit protection and power-on delay to be added to a power supply.

Power supplies with a large reservoir capacitor may draw such large currents on switch-on that problems occur, even at the primary of the mains transformer. Particularly when a toroidal mains transformer is used, it may be necessary to fit a much heavier primary fuse than is desirable for normal protection.

The current in the secondary is limited by a resistor, R1, in series with the reservoir capacitor, C1. A few seconds after switch-on, R1 is short-circuited by a relay contact. Compared with switching at the primary side, this method has the advantage that no separate supply for the relay is necessary and that this does not have to switch the 240 V mains.



Operation is fairly simple. After switch-on, C_1 is charged slowly via R_1. After a few seconds, the output voltage has risen sufficiently for the relay to be energized, which causes R_1 to be shorted. When the output of the supply is short-circuited, the output voltage drops to a level where Re_1 is de-energized. Because R_1 is then in circuit again, the short-circuit current is limited and normally the voltage regulator does not have to limit (less dissipation).

Switch S_1 enables a choice to be made between a fixed output of 12 V and one variable between 12 and 15 V.

With heavy loads it may occur that the output voltage remains too low, because of R_1, to energize the relay. In that case it will be necessary to remove the load from the supply before this can switch on.

The earth of the circuit is in a somewhat unusual place to enable IC_1 to be mounted on to the heat sink without an insulating washer (IC ground is connected to its case). For this reason, it is not permissible to use the earth for external ground connection.

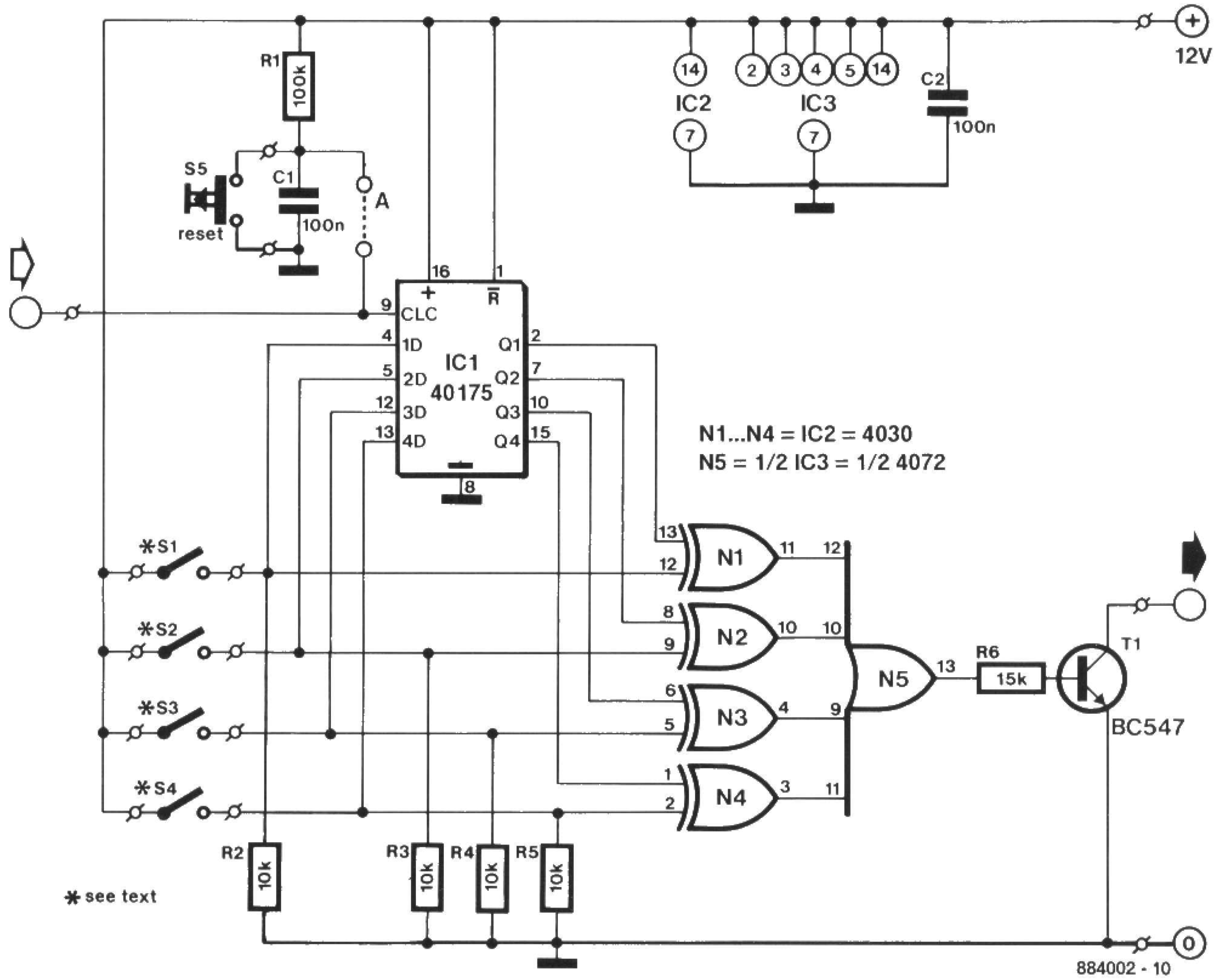
006 CAR TILT ALARM

Many cars are fitted with some sort of alarm system as protection against petty criminals and joy riders. Most of these systems rely on a door switch and one under the bonnet (to prevent interference with the battery connections to immobilize the alarm system). Such systems afford no protection whatsoever to another criminal pest: those who jack up the car and remove expensive aluminium sports wheels.

The circuit described here is an add-on to an existing alarm and energizes this when the position of the car is changed, for instance, by a jack being placed under it.

The position of the car is monitored by four mercury switches which are placed in such a way that when the car is horizontal they are open. Because a car is sometimes parked in an inclined position, which causes one or more of these switches to close, some additional circuitry is necessary.

The four D-type bistables in IC_1 determine the output state of the mercury switches. The outputs of IC_1 are connected to gates N_1 to N_4 which function as inverters when the mercury switches initially are closed (so that there is a 1 at the output of the relevant bistable). This results in the outputs of the four gates remaining 0 as long as the mercury



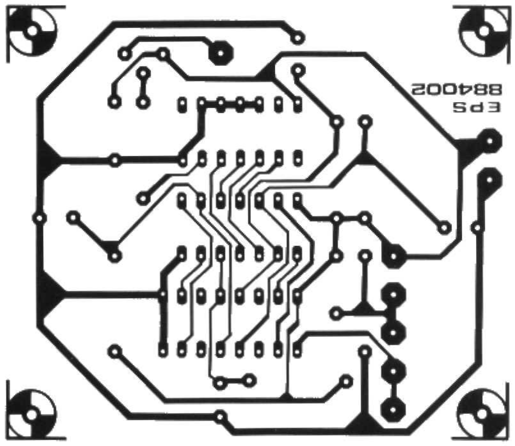
Parts list

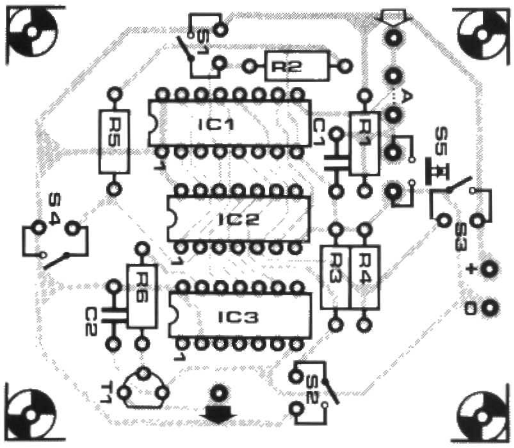
Resistors (±5%):
R1 = 100K
R2 . . . R5 incl. = 10K
R6 = 15K

Capacitors:
C1;C2 = 100n

Semiconductors:
T1 = BC547
IC1 = HEF40175BP (Philips Components)
IC2 = CD4030CN
IC3 = MC14072BCP (Motorola)

Miscellaneous:
S1 . . . S4 incl. = mercury contact.
S5 = push-to-make button.
PCB 884002 (not available ready-made through the Readers Services).



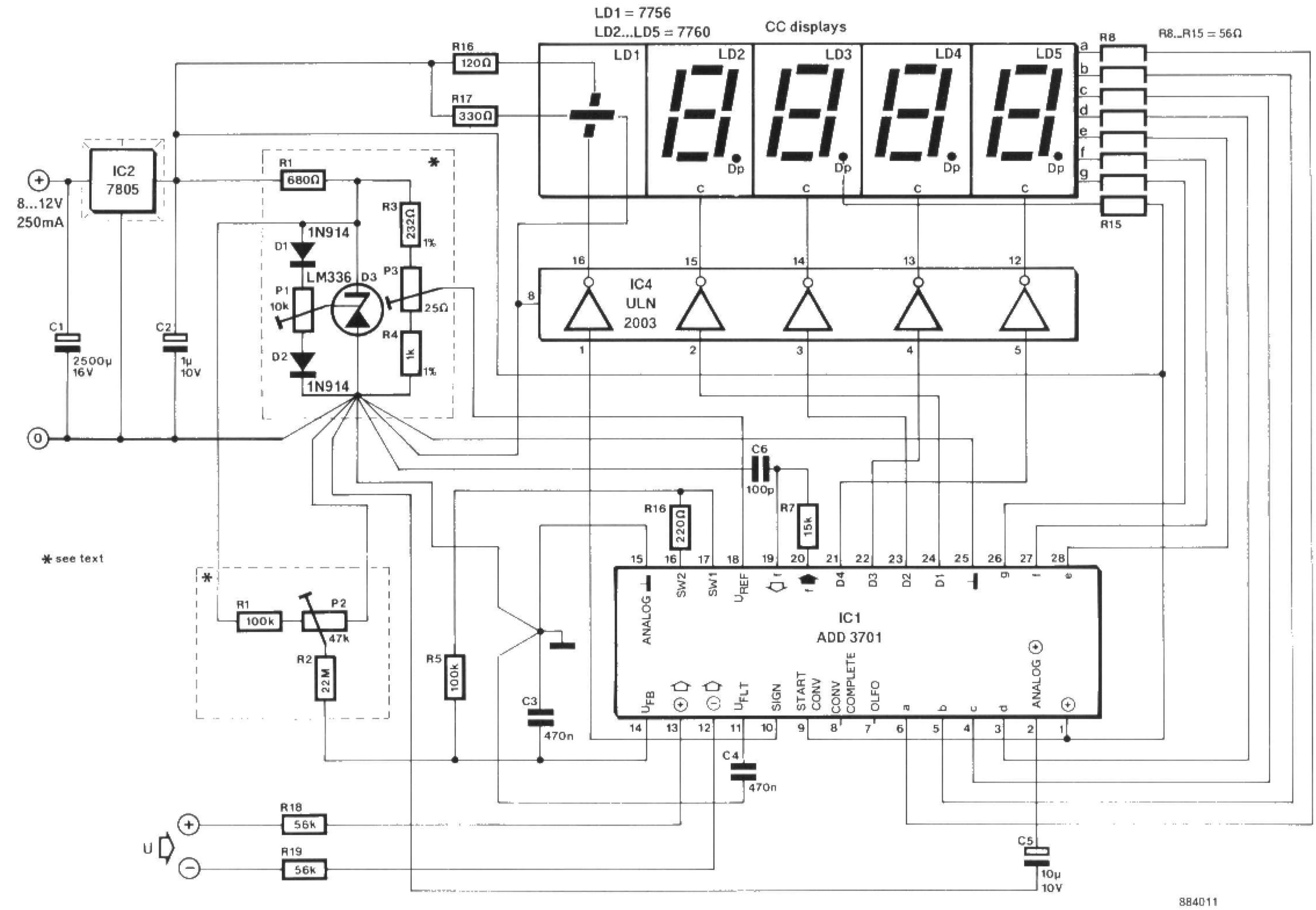


switches stay in that initial state.
If only one of the mercury switches changes state, the output of N_5 goes high and T_1 switches on. This transistor may, for instance, be connected in parallel with the door switch.
The output state of the bistables may be stored via R_1-C_1 at the moment the supply is switched on. All car alarms have a certain delay after being switched on to give the occupants time to get out of the car. If a signal is available that becomes 1 after this delay, it may also be used to store the output states in the bistables. Resistor R_1 and capacitor C_1 must then be disconnected. This second method has the advantage that if a mercury switch is just about changing state, the closing of the car doors will render it stable.
The mercury switches are mounted on the PCB together with the other components. One of the terminal wires of the switches must be kept long enough to allow the switch to be slightly tilted with respect to the board. The side of the switch in contact with the board may then be fixed into position with araldite or a similar fixative. This arrangement ensures that all switches are open when the car is horizontal.

007 3¾-DIGIT DPM

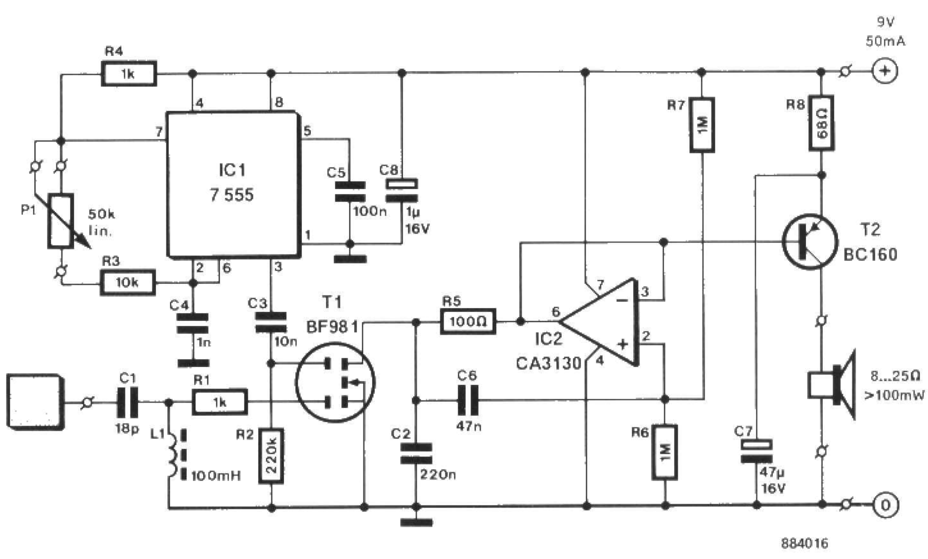
Described is a digital panel meter—DPM—which is built around a special meter-IC, Type ADD3701, and may be used for the accurate measuring of voltage from a variety of sources.
A highly stable reference voltage is provided by an LM336. A ULN2003, IC_4 is used to buffer the outputs of the ADD3701, so that the common-cathode displays can be driven direct. The ADD3701 multiplexes the displays so that the number of control lines is kept down. The current through the display segments is limited by resistors R_8 to R_{15} incl.
The oscillator that determines the conversion rate of the analogue-to-digital converter in IC_1 requires an extern RC network (R_7-C_6). Because of the need of adequate suppression of the mains frequency, the oscillator frequency must be exactly 400 Hz (it is very nearly equal to $0.6R_7C_6$). A preset potentiometer may be connected in series with R_7 to adjust the frequency accurately. At this oscillator frequency, there are about 3 conversions per second.
Another possibility of avoiding interference from the mains frequency is to use the DPM for measuring positive voltages only: LD_5 is then not required. The input voltage is applied to V_{FLT} (pin 11) via a 100 kΩ resistor. Input terminals V(+) and V(−) are not used in this case. Also, the oscillator frequency need not be exactly 400 Hz.
The DPM is calibrated by short-circuiting the input and setting P_2 to a position where the display reads 0.000. Then apply a voltage of 1.900 V to the input and adjust P_3 till the display reads 3.800. An input voltage of 1.999 V will then result in a display reading of 3.999. Take this into account if an input attenuator is contemplated.
The load presented by the input stage to a potential divider at the input is very small: typically, the input current is 1 nA (maximum 5 nA).
The (unregulated) supply should be able to provide 8 to 12 V at a current of 250 mA. The circuit, including the displays, draws about 150 mA.

(National Semiconductor Application)



008 DEFLECTION DETECTOR

Repairs to the e.h.t. section of a monitor or a television receiver always carry a certain amount of risk. It makes sense, therefore, particularly for the less experienced technician, to seek a safe way of checking the extra high tension. In all television receivers and monitors, the e.h.t. is generated in the deflection circuits. These circuits operate at about 16 kHz which generates a fairly strong magnetic field via the line transformer. It may be safely assumed that as long as the deflection circuits function correctly, the e.h.t. will also be all right. Admittedly, there is a possibility that a defect high-tension winding may be the culprit. But let's not be pessimistic.... The proposed circuit enables 'wireless' monitoring of the e.h.t. section, since it picks up all signals between about 14 kHz and 45 kHz (and their harmonics) and converts them into audio signals. The frequency of oscillator IC_1 may be varied with the aid of a potentiometer. The oscillator output is mixed with the



detected deflection signal in T_1. Since IC_2 is connected as a gyrator, filter L_1-C_1 at the drain of T_1 removes an audio signal from the mixing product. The small audio signal is amplified in T_2 to a level sufficient to drive a small loudspeaker.

The detector 'probe' is best made from a short length of insulated equipment wire, preferably, but not necessarily, connected to a small insulated metal plate. To test whether the deflection circuits operate correctly, the monitor or television receiver, as well as the test circuit, must be switched on. Then the probe should be placed in the vicinity of the line transformer and the potentiometer in the tester adjusted until a constant whistle is audible from the loudspeaker. When the monitor (TV receiver) is switched off, this whistle should disappear. If this happens, the deflection, and therefore almost certainly the e.h.t., will be all right.

009 SIMPLE TRANSISTOR TESTER

While experimenting with electronic circuits, it will often be necessary to rapidly test bipolar transistors and FETs before they are fitted in the circuit, or when they have been removed from the circuit when a malfunction is suspected. More specifically, constructors will need to know whether a transistor of known type and make is sound or not, and also whether an unknown device is a particular type of FET, or a bipolar transistor (PNP or NPN).

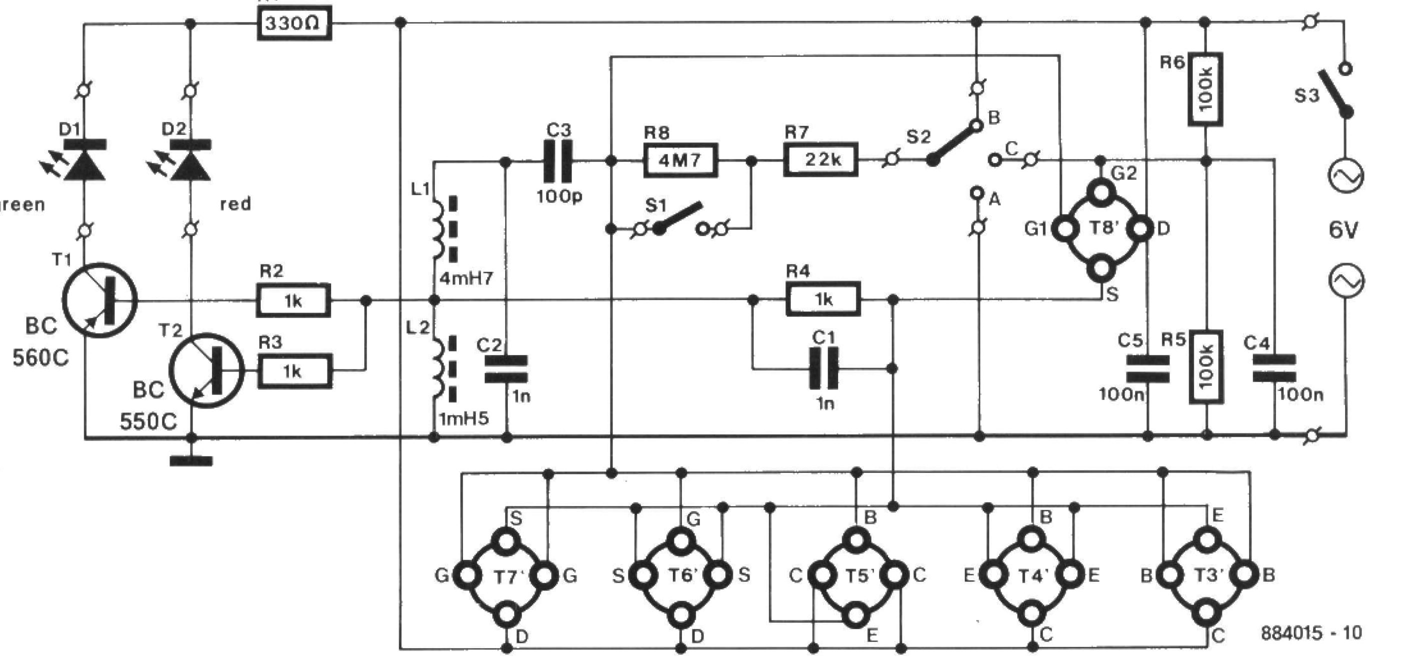
This tester can be built from parts found in the junk-box. When the transistor under test (TUT) is OK. and correctly connected, the circuit will oscillate during half the period time of the alternating supply voltage (50 or 60 Hz). Red LED D_2 lights when the TUT is OK and of the NPN type. The function of green LED D_1 is similar for PNP TUTs. The TUT OK/not OK indication is obtained with S_2 set to the centre position, and S_1 opened as shown in the circuit diagram.

The LEDs will indicate that the oscilator amplitude is significantly reduced, or nought, when S_1 is closed with a bipolar TUT mounted. Correctly operating FETs produce oscillation irrespective of the position of S_1. Only J-FETs and dual-gate MOSFETs produce oscillation when S_2 is set to positions A and C.

The accompanying table should speak for itself. Note that S_3 must be opened and closed after each change in the position of S_2.

Finally, the tester is preferably fed from a 6 VAC mains adapter.

1



Parts list

Resistors (±5%):
R_1 = 330R
R_2;R_3;R_4 = 1K0
R_5;R_6 = 100K
R_8 = 4M7

Capacitors:
C_1;C_2 = 1n0
C_3 = 100p
C_4;C_5 = 100n

Inductors:
L_1 = 4mH7 radial choke, e.g. Toko Type 181LY472 (Cirkit).
L_2 = 1mH5 radial choke, e.g. Toko Type 181LY152 (Cirkit).

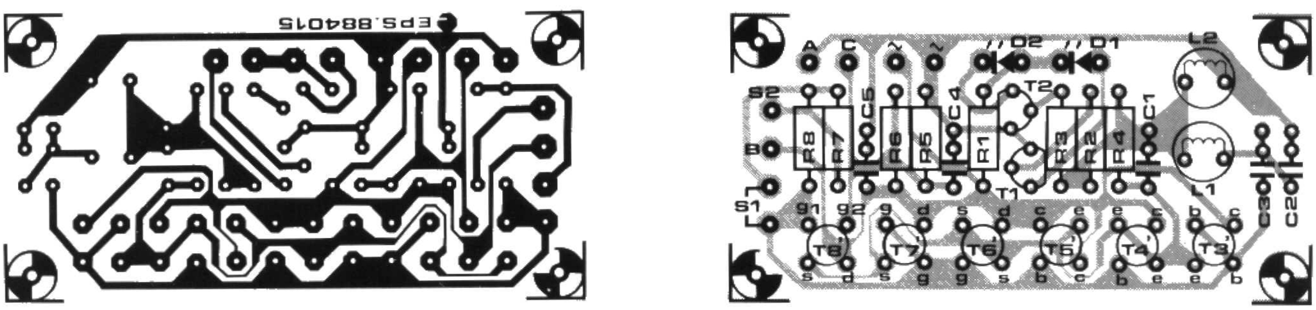
Semiconductors:
D_1 = green LED
D_2 = red LED
T_1 = BC560C
T_2 = BC550C

Miscellaneous:
S_1 = miniature SPST switch.
S_2 = 3-way rotary switch.
6 off 4-pin transistor sockets.
PCB Type 884051 (see Readers Services page).

SWITCH / TUT	S_1	S_2 (B o, A o, o C)	
BIPOLAR	● ○ ○	● ●	
J-FET	x	● ● ●	
DG-MOSFET	x	○ ● ○ (U_{g2} = ½ U_d)	● ● ● (g_1 to g_2)
ENHANCEMENT (MOS) FET	x	● ○ ●	

● = oscillation
○ = no oscillation
x = irrelevant

2



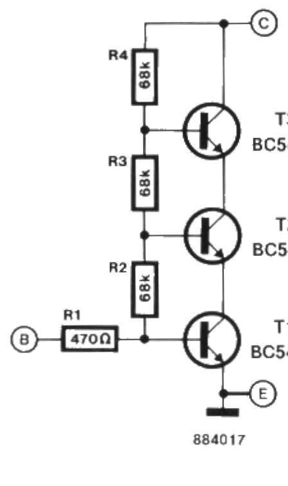
010 HIGH-VOLTAGE BC547

It is sometimes desired to use a BC547 at rather higher voltages than permitted according to the data book. Yet, it can be done by connecting a number of them in series as shown in the accompanying diagram.

The set-up has a few, small, disadvantages: there is a constant leakage current through the series resistors and the saturation voltage is rather higher.

Where these disadvantages are of little or no consequence, the circuit shown here can be used with voltages up to about 100 V.

Assume that a voltage of 100 V has to be switched and that the maximum current is 2 mA. If the current amplification is



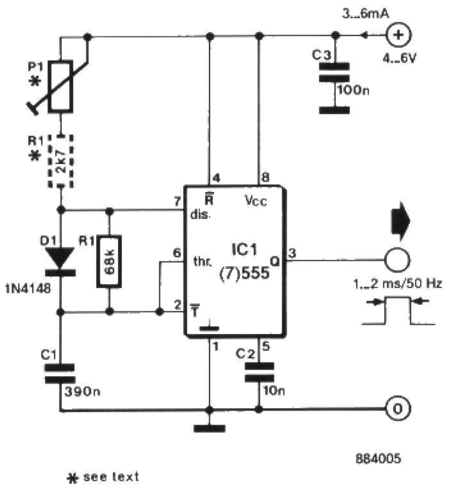
200, the base current will be 10 μA. Transistor T_3 will then switch on as soon as the p.d. across R_4 is 0.68 V. The base current of T_2 also flows through R_4, so that the drop across this resistor rises to 1.36 V.

The current that switches T_1 flows through R_1, so that it does not cause an additional p.d. across the potential divider. There is, of course, the usual saturation voltage of about 0.2 V across T_1. The total drop across the divider is then $3(10^{-5} \times 68 \times 10^3) + 0.2 = 2.2$ V.

Increasing the resistor values to 270 k raises the saturation voltage to 8.3 V. The leakage current is then much smaller.

011 SERVO-PULSE GENERATOR

Circuits for the generation of control pulses for servo apparatus remain popular, which seems a good enough reason to present another one.
The popularity of servo control is enhanced by the low price of servo motors, and the fact that they can be used for a variety of applications. The present design is geared to stand-alone use of the servo.
Simplicity of the circuit was the first design consideration, and it seemed reasonable, therefore, to base it on the well-known 555 IC. Unfortunately, this chip has the property, in its standard configuration, of producting pulse trains with a duty factor* of 50% or greater. This is so, because the charging time constant is always greater than the discharge one, since during charging the discharge resistance is in series with the charge resistance.
Servos, on the other hand, require pulse trains with duty factors well below 50%. Ideally, the pulses should have a width of 1 – 2 ms, and the pulse repetition frequency – prf – should be about 50 Hz. This gives a duty factor of 5 – 10%.

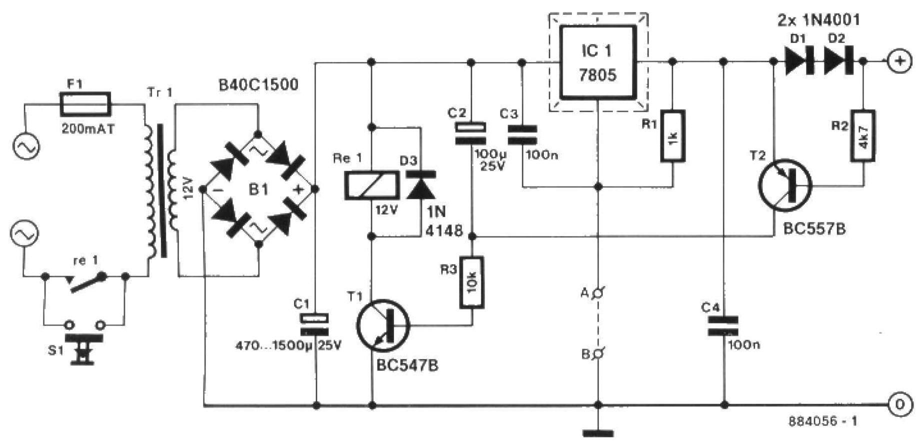


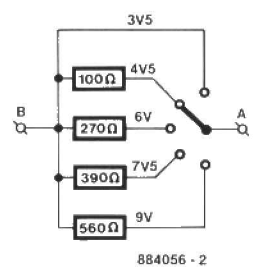
This problem may be resolved by inverting the output signal of the 555 with the aid of a transistor and two resistors, but this was considered extravagant. All it needs is an extra diode and relocating the discharge resistance. The charging time, and therefore the length of time that the output is logic high, is now determined by P_1, R_1, and the discharge time through R_2.
The component values in the circuit have been chosen in a manner that causes the pulse width to change from 1 ms to 2 ms when the resistance between the positive line and the anode of D_1 is increased from 2k7 to 5k4. This reduction in resistance is brought about by a 75° shift of P_1 (normal joystick travel), if this potentiometer has a value of 10k. This potentiometer must be set to a position where its resistance is 4kl when the joystick is at centre position. Resistor R_1 should then be replaced by a wire link.
It is possible to use the normal 270° travel of the potentiometer, which should then have a value of 2k7. Resistor R_1 must then be used as shown.

*The duty factor of a pulse train is the ratio of the average pulse width to the average pulse spacing of pulses in the train.

012 SELF-SWITCHING POWER SUPPLY

The proposed power supply switches itself off when no current is drawn by the load. How this is done is shown in the circuit diagram, Fig. 1.
When a load current flows, the p.d. across D_1 is sufficient to cause D_2 and T_2 to conduct. T_1 is then switched on and the relay is energized. When the load current ceases, T_2 switches off. The base current of T_1 will then charge C_2 so that after a few seconds the relay is de-energized. The relay contact, re_1, will then switch off the mains at the primary of the transformer. The supply is switched on again by reconnecting the load and pressing S_1 briefly.
The output voltage depends on the resistance between A and B. A wire link there results in an output voltage of about 3.5 V. For each 100R increase, the output voltage will rise by about 1 V (the current from the regulator to ground is a nearly constant 10 mA). This makes it possible to obtain a variable output voltage with the aid of some resistors and a rotary switch as shown in Fig. 2.
The relay, Re_1, should be of a type that is suitable for switching mains voltages.
The a.c. rating of the secondary of Tr_1 must be about 1.5 times as high as the desired d.c. output current. The output current should not exceed 1A; if that magnitude of current is drawn regularly, it is recommended to increase C_1 to 1500 μF.
The delay in switch-off may be extended by increasing the value of C_2.
The heat sink of IC_1 should be in accordance with the output current.





013 LCD FOR Z80-DRIVEN COMPUTERS

There is a growing tendency to use liquid-crystal displays (LCD) as the screen of computer monitors. Such displays may also be used where the normal monitor is too large or draws too much current; they are readily available. An LCD is normally driven by a microprocessor: in the proposed circuit by a Z80.

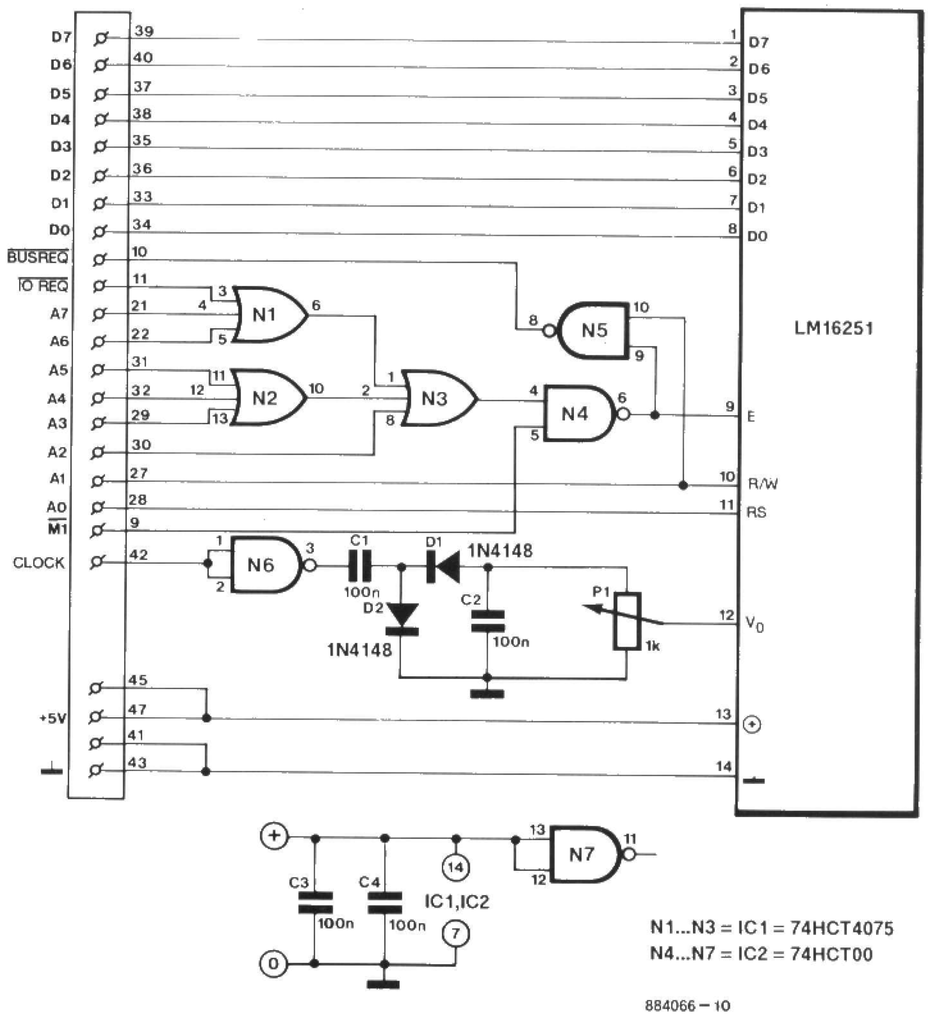
The display in the proposed circuit is a Sharp Type LM16251: a full description of this appeared in the May 1986 issue of this magazine. It is located in the I/O region, addresses 0 to 3, of the processor. This arrangement enables the circuit also to be used in combination with the 32-bit I/O and timer cartridge described in the January 1987 issue of this magazine. This cartridge does not use the lowest four addresses (choose address 0 for the cartridge so that an additional I/O region of 0 to 15 is obtained).

The address coding is effected by gates N_1 to N_4. When A_2 to A_7 are, and IO REQ becomes low, the output of N_3 goes low. If $\overline{MI}$ is high (no interrupt demanded), N_4 outputs a 1 and an enable signal is given to the display.

Depending on the logic levels at inputs R/W and RS, data is transmitted or received. The RD and WR outputs of the Z80 are not used, because the R/W and RS signals of the LM16251 must be stable not later than 140 ns before the E input goes high. If the RD or WR signals of the processor were used, the E input of the display would be accessed together with the other signals and that is not permitted.

By using an address line, the timing is arranged by that of the Z80, because the address bus must be stable not later than 320 ns (180 ns for Z80A) before an IO REQ signal is generated. Owners of a Z80B-driven computer might have some problems here because the time delay is then only 110 ns. Note that MSX computers invariably use a Z80A.

The negative voltage for the contrast setting (P_1) of the display is provided by



N_6. Note that some types of display need a positive voltage for the contrast setting. Wire link 'a' provides a negative supply, and 'b' a positive one. Link 'a' is required for the LM16251. If another type of display is used, make sure that the pin numbering is the same as shown in the diagram.

Gate N_5 serves to render the $\overline{BUSDIR}$ line low at an I/O read command in MSX systems. In other systems, this gate is not required.

014 CAR INTERIOR LIGHT DELAY

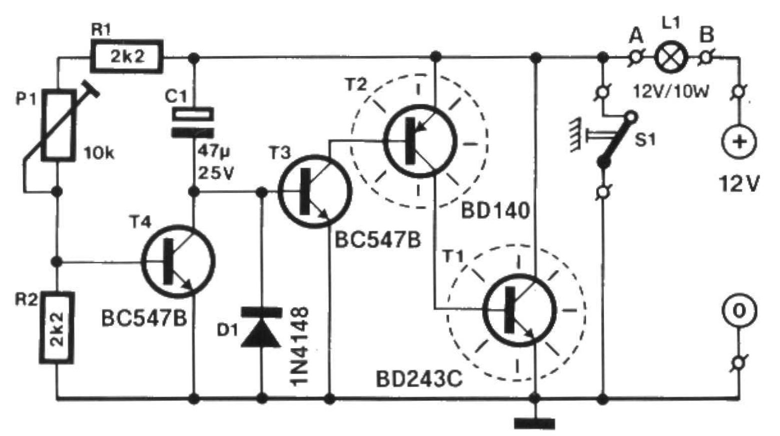
It's dark and it's raining cats and dogs. You rush to your car, open the door and quickly close it behind you again. Then you sit there fumbling for the ignition lock. Solution? Add the following circuit, which will keep your car's interior light on for a little while after the door is closed.

The circuit is connected across the switch in the door post. These switches are removed quite easily.

In the circuit diagram, S_1 is the switch in the car's doorpost and L_1 is the interior light. As long as the door is open, S_1 is closed and the light is on. When the door is closed, S_1 opens and the light goes out. The full 12 V from the car battery is then present across the switch.

The circuit detects when the voltage across S_1 begins to rise. Transistor T_3, and consequently T_1 and T_2, is then switched on. This results in the voltage across S_1 rising to about 1 V, after which it can increase only very slowly. This means that the interior light stays on, although its brightness will slowly decrease.

At a certain level of potential across S_1, transistor T_4 switches on, which results in the drive to T_3 becoming zero, and T_3,



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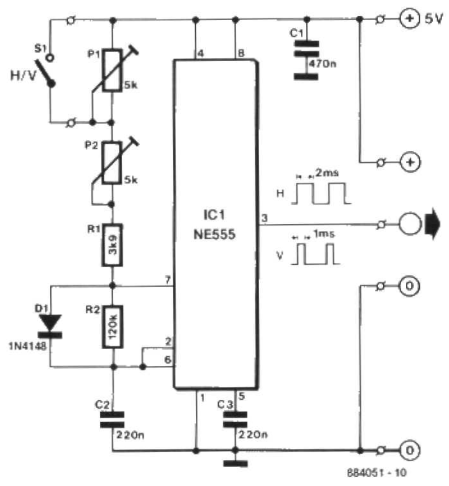
T_2, and T_1 switch off. The interior light will then go out very quickly.
The delay in the light going out after the car door is closed is preset by P_1, although it is also affected by the value of C_1. The larger this value, the longer the delay and the smaller the variation in the brightness of L_1. After the light has gone out, the circuit no longer draws current.

015 POLAROTOR CONTROL

The polarization of satellite TV signals is defined as horizontal (H) or vertical (V) with respect to the equator below the subsatellite point, and not, as is often wrongly assumed, with respect to the horizon on earth. Depending on the location of the receiving system on earth and the satellite's geosynchronous position, a horizontally polarized signal may have some offset with respect to the horizon. As a rule of thumb, the lower the dish elevation for a particular satellite, the greater this polarization offset angle. The difference between horizontal and vertical is, however, always 90°.
Most commercially available polarization rotation units (*polarotors*) used for selecting between horizontally and vertically polarized transponders on board a TV satellite incorporate a small servo motor whose direction of travel is controlled automatically by the channel selection circuitry in the indoor unit, or simply by a switch. The servo motor rotates an angled probe fitted in a PTFE bush in the waveguide flange that is secured onto the feed horn. This probe can be rotated over 90°, and re-transmits the received 11 GHz satellite signal by means of a ¼λ probe fitted vertically in the waveguide that connects to the LNB.
The polarotor assembly is fitted permanently between the feed horn and the LNB input, and is connected to the indoor unit via a length of 3-wire cable, which runs in parallel with the downlead coax. A polarization selection switch, S_3, is provided on the Indoor Unit for Satellite TV Reception (1), but not the accompanying driver circuit, which is given here.
The polarotor control is an astable multivibrator that determines the direction of travel of the servo motor by supplying output pulses with a duration of 1 ms (V) or 2 ms (H) (typical values). When horizontal/vertical (H/V) switch S_3 is closed, P_1 is short-circuited, so that IC_1 supplies pulses with a duration of 1 ms. In the polarotor assembly, a com-



bination of a potentiometer coupled to the motor spindle and an electronic circuit is used for comparing the duration of the received control pulses with that of the internally generated spindle positioning pulses, and actuates the motor until the pulses are of equal duration. The microwave probe in the feed horn waveguide is then positioned vertically. Similarly, when S_3 is opened, P_1 is included in the R-C timing circuit of IC_1. Due to the higher total resistance, IC_1 supplies pulses with a duration of about 2 ms, so that the waveguide probe is rotated over 90° for reception of horizontally polarized signals.
The control circuit and the servo motor are powered from a regulated 5 V supply, which is simple to construct around a Type 7805 3-pin integrated regulator. In the case of the above mentioned Indoor Unit, the input of the 7805 can be connected to the input of IC_7 (Type 7812 on the vision/sound/PSU board). Due care should be taken, however, not to overload the mains transformer, Tr_1, or optional series resistor R_x, since the maximum current consumption of a blocked polarotor motor is typically about 300 mA. In some cases, it may be necessary to fit a relatively large electrolytic decoupling capacitor direct across the supply terminals of the servo motor. The value of this capacitor depends on the actual current consumption of the motor, but 470 µF should work satisfactorily in most cases. It is recommended to use fairly stout wire for connecting the polarotor to the control circuit.
The circuit is simple to set up: connect an oscilloscope to the pulse output line, and adjust P_1 and P_2 for correct duration of the rectangular output pulses (note that the settings interact). Open the available polarotor to check that the travel of the probe covers the full range of 90°. In the absence of an oscilloscope, P_1 and P_2 are adjusted until the servo motor works reliably over the full range in both directions of travel. Polarization offset correction can be achieved by adjusting the presets accordingly. Continuous adjustment of the probe position (*skew*) for satellite reception experiments can be achieved by using potentiometers instead of presets in positions P_1 and P_2. Current consumption of the control circuit is about 7 mA.

(1) Indoor Unit for Satellite TV Reception. *Elektor Electronics* October & November 1986, January 1987.

016 PROGRAMMABLE VOLTAGE SOURCE

A number of appliances, such as an EPROM programmer, require a supply voltage that can be switched to a variety of levels. The proposed circuit enables the user to do so between 5 V and 21 V. As soon as the switching transistor conducts, R_3 is connected in parallel with R_2. This lowers the total resistance between the 'adj' pin of the LM317 and earth, and consequently the output voltage.

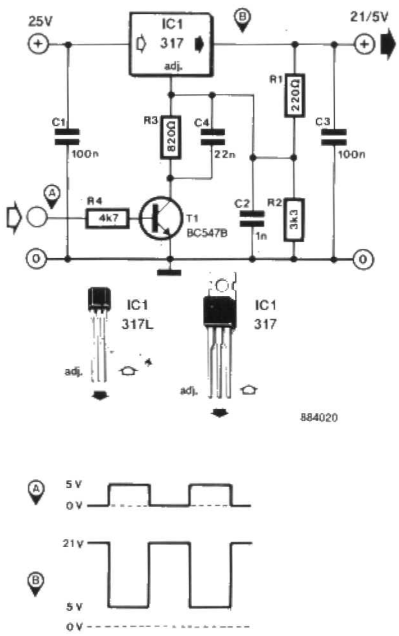
It is possible to add a number of switching transistors and associated resistors and capacitor to the circuit to increase the number of available output voltage levels.

The level of the output voltage depends on the ratio between R_1 and the resulting value of R_2 in parallel with R_3. The p.d. across R_1 is always 1.2 V. Thus,

$$U_o = [1 + \frac{R_2/R_3}{R_1}] \text{ volts}$$

Capacitors C_1 and C_3 serve to optimize the switching behaviour of the circuit. The value of these components has to be established with the aid of a squarewave generator and an oscilloscope. The effect of these capacitors on the output voltage is shown in the photograph.

1



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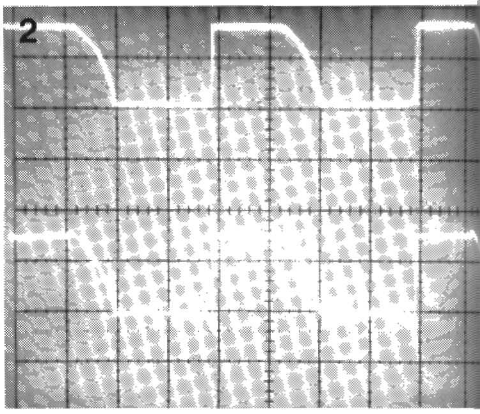

An additional advantage of the use of an integrated voltage regulator is that this affords a means of current limiting. If, for instance, the 'L' type of this IC is used, current limiting starts at about 100 mA. This magnitude of current will be more than adequate for most EPROMs.

Finally, it is possible to replace T_1 and R_4 by a high-voltage open-collector TTL gate, such as provided by the 7407.

017 AUTOMATIC VOLUME CONTROL

The proposed AVC gives weaker components of the input signal extra amplification while ensuring that this dynamic compression is not disconcerting. It therefore eliminates those annoying differences in loudness between speech and music on radio and television.

The principle of the circuit is fairly simple. Field-effect transistor T_1 is used as a variable resistance. The value of this resistance, $r_{DS(on)}$, can vary from infinity to about 150R. It is in parallel with R_3 and, in conjunction with R_4, determines the gain of A_1. Without the effect of the FET, the gain of A_1 is about 20 dB.

Opamp A_2 is connected as a straightforward amplifier, whose gain may be varied by P_1. The negative part of the output signal of A_2 is connected to the gate of T_1 via a rectifier formed by D_1, C_1, R_7, and R_8. Resistor R_8 ensures that the switching of T_1 happens gradually. This means that it takes a short time before T_1 operates; in other words, momentary differences in input level do not affect the overall gain. The reduction in gain also takes place gradually, because C_1 has to discharge via R_7.

Because the resistance of T_1 is influenced by the drain-source voltage, U_{DS},

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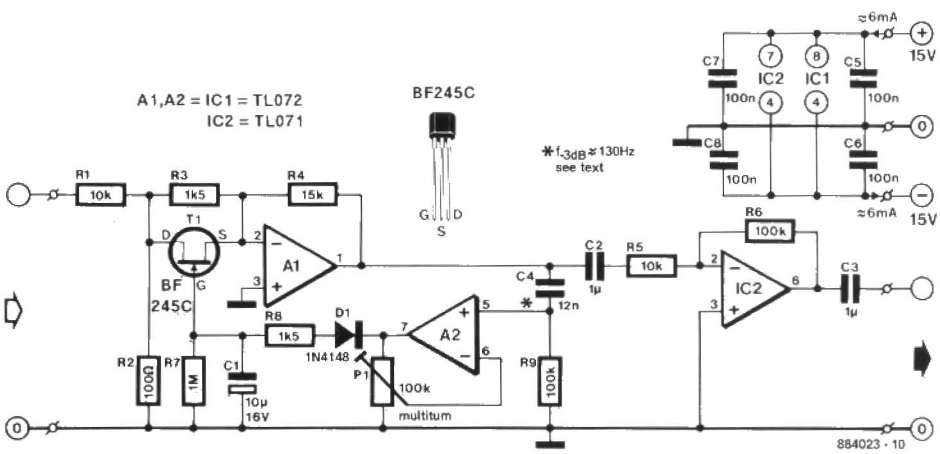


the signal level has to be kept as low as possible (thanks to the use of opamps, there is no direct voltage across the drain-source junction). An attenuator, R_1-R_2, which gives an attenuation of 40 dB, is therefore provided at the input. This enables signals of up to 1 V r.m.s. to be processed with a distortion of not greater than 0.6%. With an input of 1 V r.m.s., the signal-to-noise ratio is about 70 dB.

The amplification in A_1 and A_2 compensates the losses in the attenuator: the total gain of the circuit, with T_1 switched

off, is 0 dB.
Network R_9-C_4 is a high-pass filter which ensures that strong bass signals do not affect the control function to much extent. The cross-over point may be altered to personal taste.
Signals at a level below that set by P_1 are amplified by a factor of up to 6.9 (gain = 17 dB). Fig. 2 shows the relation between input and output levels.
The circuit needs a supply voltage of ± 15 V and draws a current of about 6 mA.

Parts list

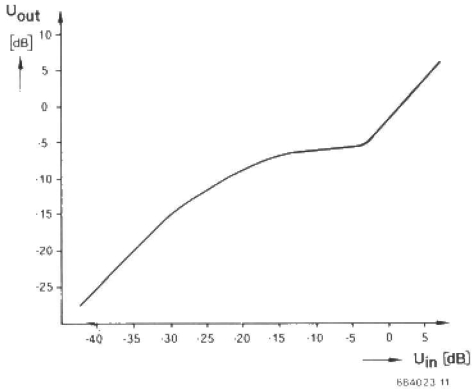
Resistors (±5%):
R_1;R_5 = 10K
R_2 = 100R
R_3;R_8 = 1K5
R_4 = 15K
R_6;R_9 = 100K
R_7 = 1M0
P_1 = 100K multiturn preset

Capacitors:
C_1 = 10µ; 16 V
C_2;C_3 = 1µ0; MKT
C_4 = 12n
C_5...C_8 incl. = 100n

Semiconductors:
D_1 = 1N4148
T_1 = BF245C
IC_1 = TL072
IC_2 = TL071

Miscellaneous:
PCB Type 884023 (not available ready-made through the Readers Services).

2



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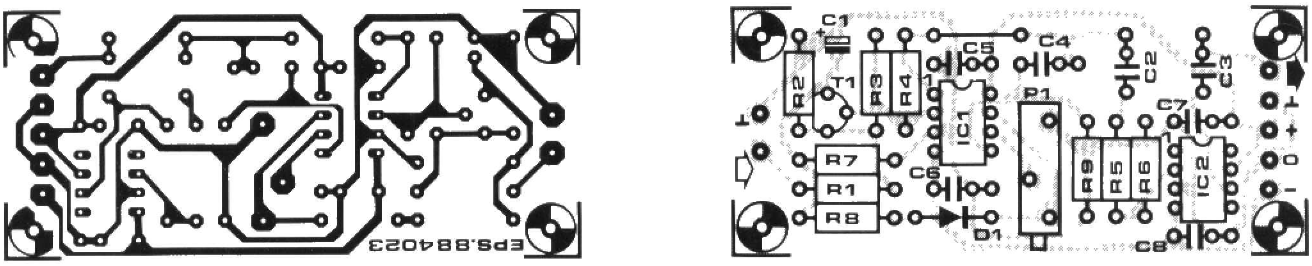

018 SAMPLE & HOLD FOR ANALOGUE SIGNALS

Conventional analogue sample and hold circuits are notorious for their tendency to drift, a phenomenon unknown in digital memories. It is, therefore, interesting to study the use of a digital memory element for storing an analogue signal.
The present circuit is based on intermediate storage of digitized analogue information, and therefore requires an analogue-to-digital converter (ADC) at the input, and a digital-to-analogue converter (DAC) at the output. Unfortunately, DACs and ADCs are typically expensive components, and the present circuit is therefore set up with a DAC only, driven by an up/down counter—see Fig. 1. The counter is essentially an ADC, since the output voltage of the R-2R based DAC is continuously compared to the input voltage with the aid of a window comparator. The error signal produced by the comparator arranges for the counter to count up or down, depending on the magnitude of the difference between the input and output voltage. The up/down counter is corrected until the input and output voltage are equal. The digitized result of the A-D conversion is available at the counter outputs.
The extensions for converting the basic set-up into a sample & hold circuit are relatively simple. The current count is retained by activating the $\overline{HOLD}$ input, which enables halting the U/D counter. Evidently, the counter state is not subject to drift, so that the analogue output

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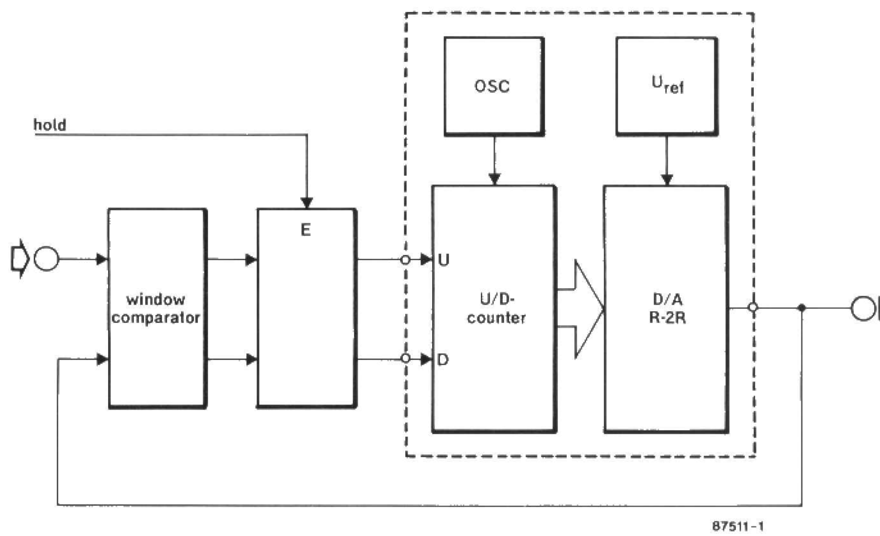

signal is available unaffected for as long as the circuit is powered. The converter used here is the Type ZN435 ADC/DAC from Ferranti. This chip contains everything shown in the dashed box of Fig. 1. With reference to the practical circuit diagram, Fig. 2, the internal voltage reference and the oscillator are adjusted with R_1-C_1 and R_2-C_2 respectively. The latter are dimensioned for 400 kHz, i.e., nearly the maximum oscillator operating frequency. The internal counter is controlled via inputs $\overline{up}$, $\overline{down}$ and mode. The logic level applied to the mode input determines whether the counter continues or halts upon reaching state 0 or the maximum value, 255. In the present application, the counter is halted. Gates N_1 and N_2 are added to enable blocking the U/D counter. Opamps A_1-A_2 form the window comparator. Current source T_1-R_7 and R_6 arrange for the toggle threshold of A_1 to be 20 mV higher than that of A_2. This off-set creates the window, or inactive span, needed to suppress oscillation of the counter's LS bit, and to prevent unwanted effects arising from the comparators' offset voltages. Decoupling capacitor C_3 is fitted for suppressing spikes that occur during state changes on the counter outputs. The

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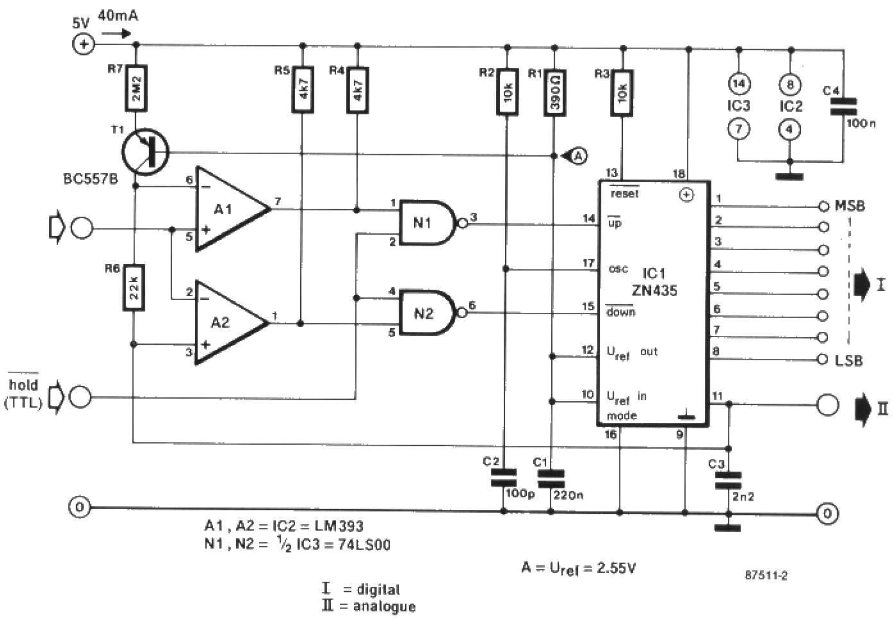


conversion time of this design is about 640 μs, as determined by the oscillator frequency (400 kHz), the resolution (8 bits) and the input voltage change (2.55 V_{pp} max.). This corresponds to a slew rate of 4 mV/μs at the input. Finally, bear in mind that the output impedance (IC_1, pin 11) is relatively high at about 4 kΩ.

019 LOW-FREQUENCY LC OSCILLATOR

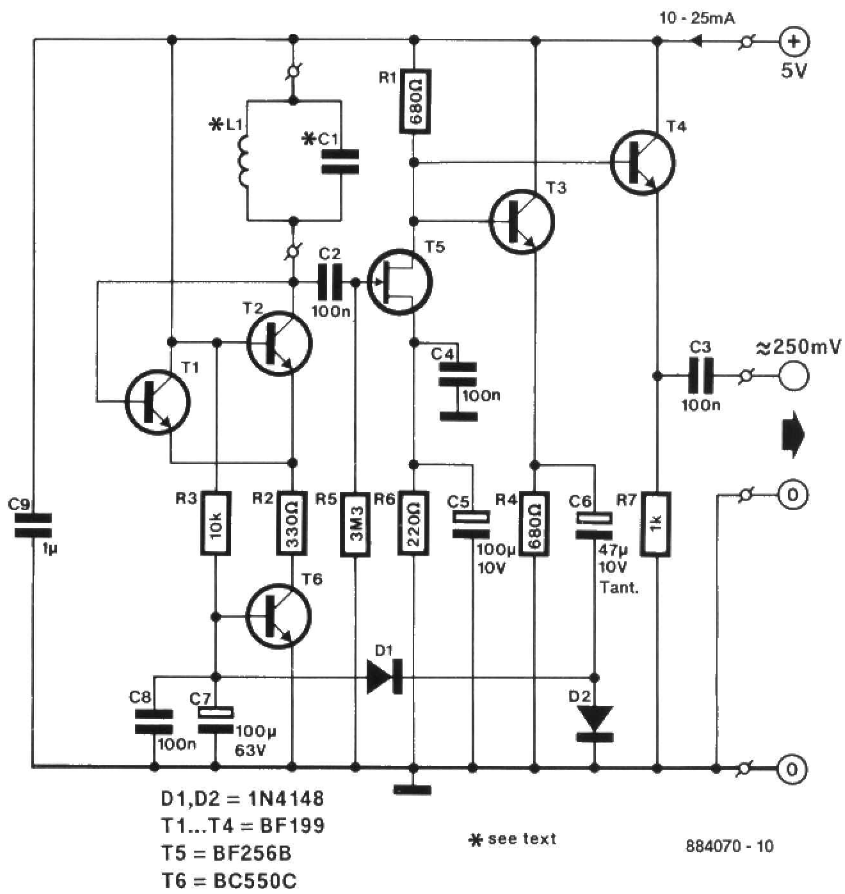
It is not always appreciated that LC circuits may be used for generating low frequencies. The proposed circuit, provided it uses good-quality components, can be used for frequencies down to 150 Hz, and possibly even slightly lower. The oscillator proper consists of T_1 and T_2 with the LC circuit connected in the collector circuit of T_2. The amplification is set with the aid of the current source around T_6.

The voltage across the tuned circuit is tapped at high impedance and amplified by T_5. The output of this FET is buffered by T_3 and then rectified by D_1—D_2. The resulting direct voltage is used for driving the current source. Since the rectified voltage still contains a ripple, a further buffer, T_4, is added at the output of the circuit.

The circuit draws a current of about 20 mA, which can rise to about 25 mA at higher frequencies. Its output impedance has been kept as low as possible to render the bandwidth of the oscillator as broad as possible.

Fairly high values of inductance may be used, provided the Q is of a reasonable value. Capacitor values may go up to 10 μF, but note that electrolytic types can not be used.

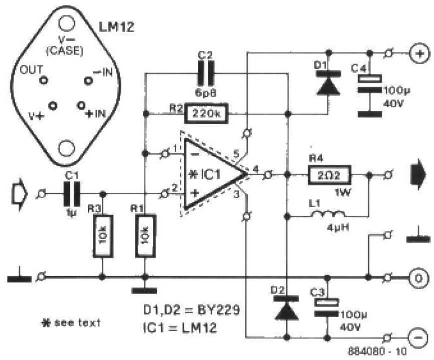
In the prototype, L_1 had a value of 150 mH and C_1 was 6μ8: the resulting frequency was 150 Hz. The oscillator generates pure sine waves up to 7 to



8 MHz and operates well up to about 30 MHz; the waveshape is then no longer a pure sine wave, however. Operation at still higher frequencies is possible, but the output level then drops from the nominal 250 mV.
The circuit may be used to measure unknown capacitors or inductors, provided the other component in the LC network is known, with the aid of the formula $f = 1/2\pi\sqrt{LC}$.

020 SINGLE-CHIP 150 W AF POWER AMPLIFIER

The Type LM12 operational amplifier from National Semiconductor has at least one remarkable characteristic: its huge output current capability of about 10 A. The chip is housed in a 4-pin TO-3 enclosure, can handle peak powers up to 800 W, and has extensive internal protective circuits to prevent damage caused by current and voltage overloading, or by overheating. Peak operating temperature of the on-chip power output transistors is measured for controlling a limiter that forms part of a so-called *dynamic safe area protection circuit*.
The power output stage is not connected to the relevant pin until the supply voltage exceeds 14 V (±7 V). Output disconnection is automatic when the chip temperature rises above 150 °C. It is possible to connect LM12s in parallel, or in a bridge configuration, for very high power applications (voltage regulators, automotive drivers, stepper motor or power servo controllers, etc.).
The present application discusses the use of the LM12 in a high-power AF amplifier.
The circuit diagram shows two clamping diodes at the chip output. These prevent the output voltage swing exceeding the supply voltage when the push-pull output stage in the chip is overdriven, and the output load is mainly inductive. The diodes also protect the chip when the output is short-circuited to the positive or negative supply rail. The Type LM12CL or LM12C may be used with supplies up to ±30 V or ±40 V respectively.
Input bias currents are compensated because the circuit is laid out for virtually equal impedance at the inverting and non-inverting input of the opamp. Input offset is 20 mV maximum. If this is considered too high, it can be cancelled completely by applying an appropriate offset compensation voltage to one of the inputs (use a well-decoupled potential divider). Output offset voltage in a number of prototypes without compensation circuitry was between 100 and 200 mV.
Half-power (−3 dB) bandwidth of the amplifier is 16 Hz to 40 kHz; distortion is approximately 0.02% at $P_o = 1$ W and $R_L = 2\ \Omega$ or $4\ \Omega$. At full drive, distortion increases to 0.05% ($U_b = \pm 30$ V; $R_L = 4\ \Omega$). Maximum current is supplied to a 2 Ω load, but distortion then increases to 0.1%.
Quiescent current of the amplifier is between 65 and 100 mA. Inductor L_1 is wound as 40 turns of 1 mm dia. enamelled copper wire on power resistor R_4. It serves mainly to ensure correct operation of the feedback amplifier with capacitive loads, such as large voice coils and loudspeaker cross-over filters.
It will be clear that the supply for the amplifier must be capable of handling the peak current requirement of the LM12. For the LM12CL, it is recommended to use a toroid mains transformer with a 2×22 V secondary winding (150 W can then be supplied to a 2 Ω load only). Depending on the application and the output power required, the transformer's secondary



should be rated between 7 and about 12 A. Smoothing capacitors in the symmetrical supply should be not smaller than 20,000 µF on each rail.
Finally, IC_1 should be bolted on to a large heat-sink, from which it is electrically insulated.

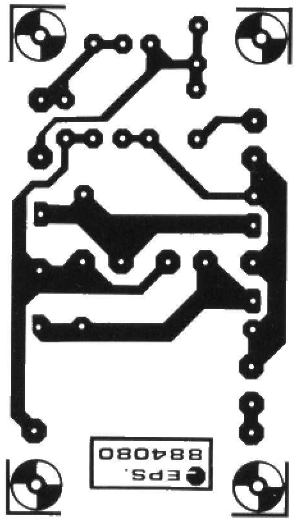
Parts list

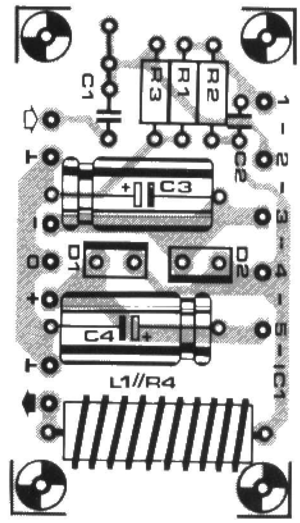
Resistors (±5%):
R1;R3 = 10K
R2 = 220K
R4 = 2R2; 4 W

Capacitors:
C1 = 1µ0; MKT
C2 = 6p8
C3;C4 = 100µ; 40 V

Semiconductors:
D1;D2 = BY229
IC1 = LM12 (National Semiconductor)

Miscellaneous:
Large heat-sink for IC1 (≤1.5 °C).
Insulating material for IC1.
PCB Type 884080 (see Readers Services page).

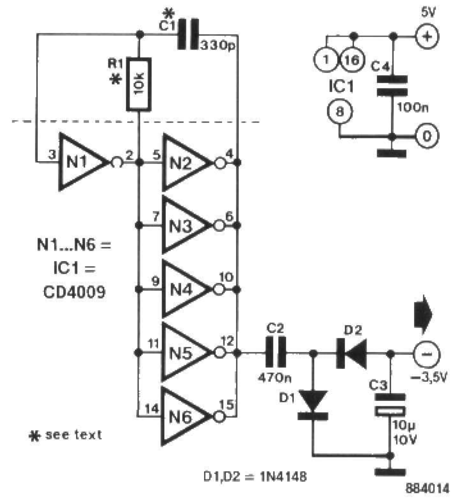




021 AUXILIARY NEGATIVE-VOLTAGE SOURCE

Many circuits require, apart from the usual positive voltage source, a negative supply from which only a small current is drawn. In such cases, a mains transformer with twin secondary winding would be a rather too costly solution.
The circuit proposed here generates a negative potential from a positive supply. This supply may provide between 5 and 15 V. If the current drawn from this supply is smaller than 1 mA, the level of the negative voltage generated lies about 1.5 V below that of the supply voltage. Thus, if the supply is 5 V, the negative potential is −3.5 V. When a current of 2 mA is drawn from the supply, the difference between the two voltages increases to about 2.5 V.
Operation of the circuit is fairly simple. Gate N_1, in conjunction with parallel-connected gates N_2 to N_6 incl., functions as a square-wave generator with buffered output. The peak-to-peak value of the square-wave voltage is, due to the use of CMOS gates, very nearly equal to the supply voltage. Rectifier D_1–D_2 ensures that the alternating voltage is converted into a steady negative one.
If a clock frequency between 10 and 50 kHz is available, this can be applied to the input of N_1. Capacitor C_1 and R_1 are then not required.

(Intersil application)



022 SINGLE-CHIP SOLID-STATE RELAY

Light-duty (25 to 600 W) solid-state relays have recently been introduced on the market by Sharp. These small and compact devices switch accurately at the zero-crossing and provide the required electrical separation. The photograph shows clearly that switching occurs exactly at the zero-crossing. This prevents switch-on currents of lamps becoming large and so extends the life of the lamp.
The breakdown voltage of the triac section is 2 kV and the pins are on a 0.1 in grid.
The relay requires an energizing current of 10 mA at 1.4 V, but with inductive loads about 25 mA is necessary.
The additional components shown in the diagram make the relay more universally usable. Diode D_1 prevents the IC being damaged if the input is connected incorrectly. Transistor T_1 sets the trigger current to precisely 10 mA. The RC network at the output protects the triac from sharp voltage peaks.
The IC may be used without heat sink to switch currents up to 1 A. For switching larger currents, up to a maximum of 3 A, a 2mm thick 100×100 mm heat sink should be used.

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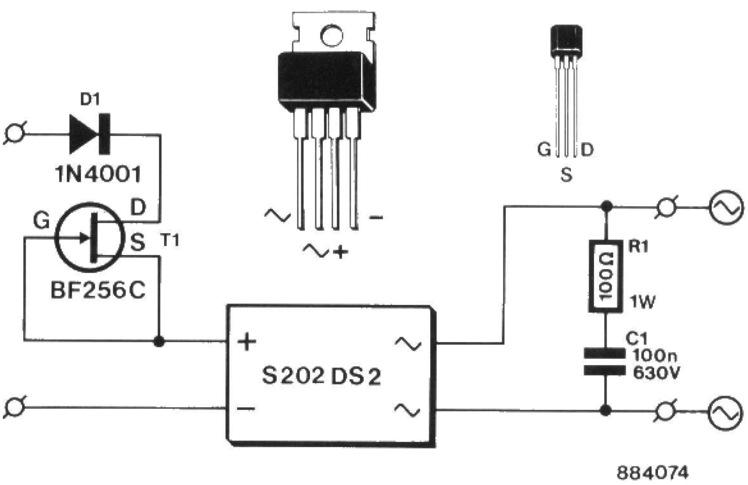

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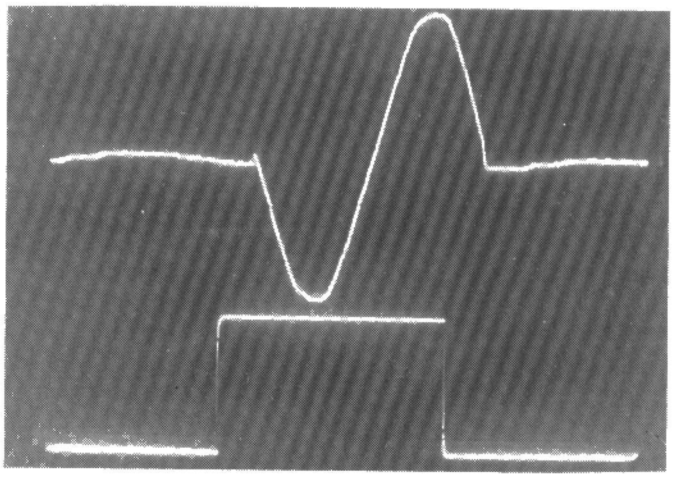


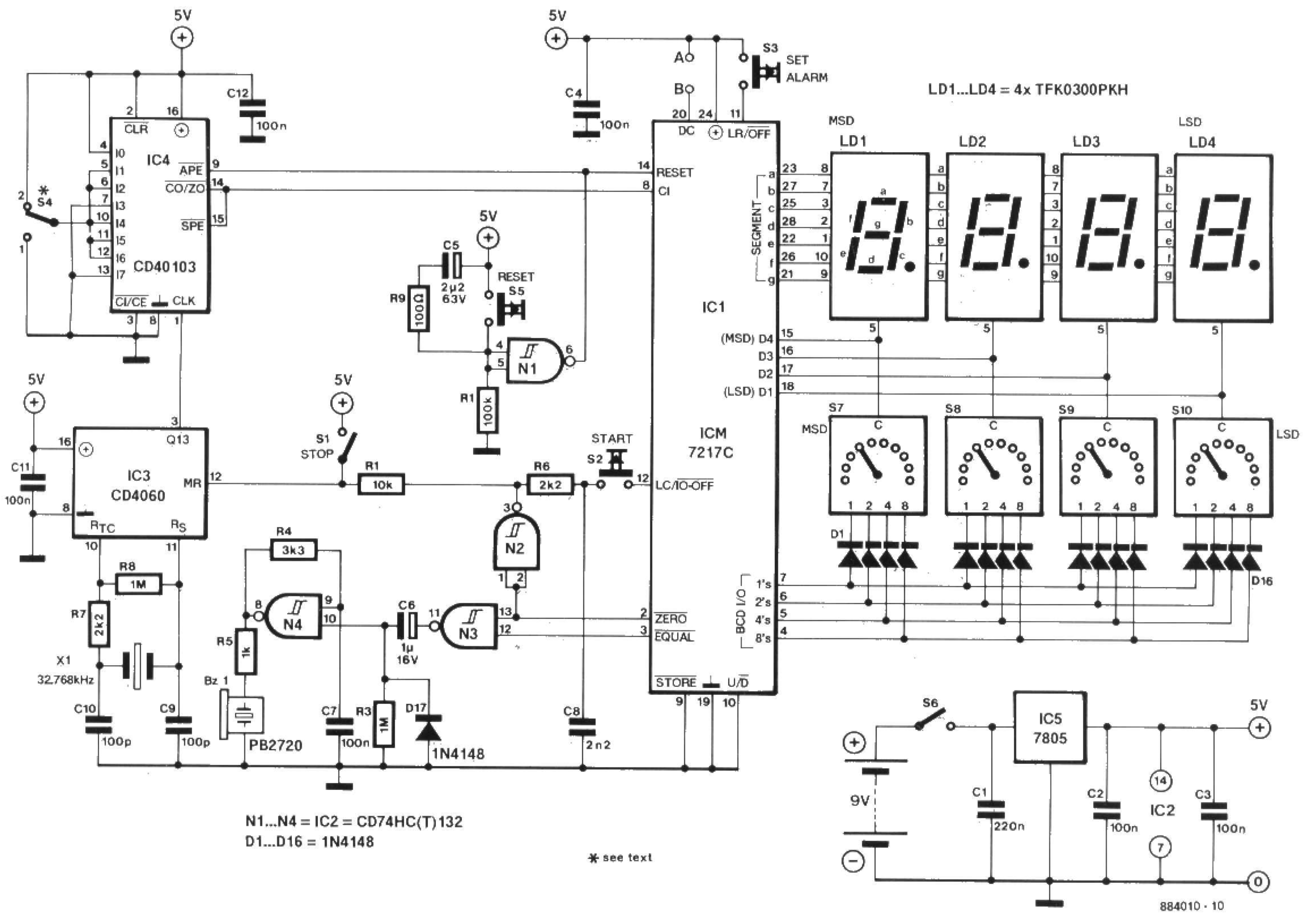
023

TIMER

This timer can be set to count a maximum of 60 hours. It also allows an interval to be set. When this interval is reached, a buzzer sounds.
The larger part of the circuit is contained in an Intersil Type ICM7217 four-digit CMOS up/down counter and display.
Circuit IC_3 is the clock that generates a square wave with a period of 1 s. The clock signal is available at pin 3 (Q13). The clock signal may be divided by 60 in IC_4 if it is required to time more than 1 hour.
When S_6 is closed, the supply is switched on and IC_1 is reset via R_9 and C_5. The position of S_4 determines whether minutes or seconds are counted: maximum 59 h 59 min (pos 2) or 59 min 59 sec (pos 1).

If, for instance, a total time of 35 min with an interval at 20 min is to be counted, S_4 is set to position 1. Thumb wheel switches S_7 to S_{10} are then set for a display reading of 20.00. Briefly pressing S_3 stores this setting in the memory of IC_1. Then S_7 to S_{10} are set for a display reading of 35.00. During these settings, S_1 should be open. Pressing S_2 causes the ICM7217C to count down from 35.00. When display reading 18.00 is reached, the buzzer briefly sounds (energized via N_3 and N_4). The timer may then be stopped by closing S_1. When S_1 is opened again, the timer restarts the down count to 00.00. When that reading is reached, the buzzer sounds briefly again. Note that at any time during the count down the timer may be stopped by closing S_1.

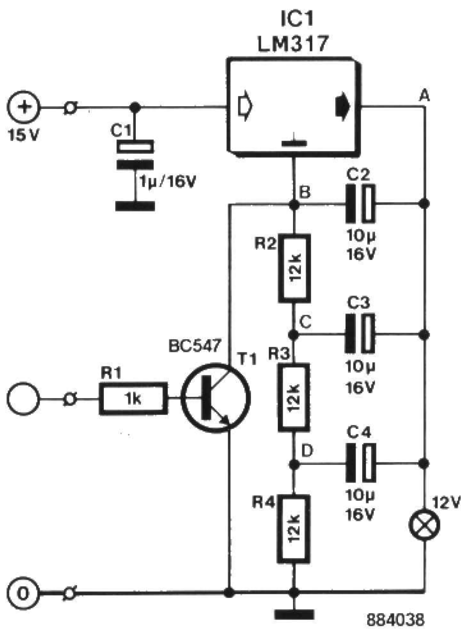
The timer is reset with S_5; when that happens, the buzzer sounds briefly and the display reads 00.00. The set count down period of 35 min is, however, retained in the memory until a new period is programmed.
The current drawn by the timer, including the displays, is about 100 mA. If a battery supply is used, it is possible to switch off the displays when the timer is counting by adding a switch (with single break contact) between points A and B. This switch enables the display to be read briefly. With the displays switched off, the current drawn is of the order of only 4 mA.
Do not set the thumb wheel switches to readings greater than 59.59, because the timer will then no longer count correctly.

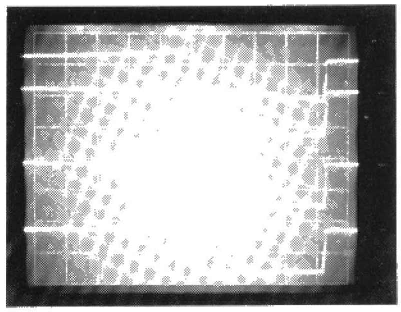


024 FLASHING LIGHT

This is a rather unusual application of the Type 317 voltage regulator. With only a handful of external components, it can be used for flashing a small 12 V lamp. The output voltage is not stabilized by the circuit: it is simply a few volts lower than the input voltage. The 317 is capable of delivering more than 1 A.
The circuit automatically limits the switch-on current, so that lamp life is considerably extended. The waveforms at the four major points in the circuit are shown in the accompanying photograph.
The component values given result in a flash frequency of about 4 Hz. Flashing can be stopped by driving T_1 with a voltage of more than +1 V.

Source: *Lambda Power Supply Handbook*





025 AMPLITUDE-MODULATED CALIBRATION GENERATOR

A calibration generator is used for quickly checking receiver operation. The design shown here generates RF signals (*markers*) at 1 MHz intervals over a frequency extending up to about 2 GHz. These signals can be amplitude-modulated by driving T_4 with a sine wave generator.
A stable 2 MHz oscillator is set up around X_1 and T_1. MOSFET T_2 functions as a digital buffer for clocking bistable/divider FF_1. Pulses at the output of FF_2 have a frequency of 1 MHz and a width of only 12 ns, which is obtained by FF_2 clearing itself after output $\overline{Q}$ has gone low. The pulses drive T_3 into saturation. This SHF transistor consequently produces a wide spectrum of harmonics, and its class C setting causes it to function as a frequency multiplier. The collector current can be modulated via series transistor T_4. Since the two sidebands generated in the process of amplitude modulation are offset from the carrier by the modulation frequency, AM can be used to generate signals at frequencies in between the markers. Example: modulating the calibration generator with a 204 kHz sine wave gives two additional frequencies adjacent to the marker at, say, 1120 MHz: $1120-0.204=1119.796$ MHz and $1120+0.204=1120.204$ MHz. Hence, a continuous tuning range from 1 MHz to 2 GHz is obtained when the sine wave generator output frequency is adjustable between 500 kHz and 1 MHz.
The measured amplitudes of four markers produced by the calibration generator show that available output levels fall with increasing frequency:

f=100 MHz: $P_o=-25$ dBm
f=400 MHz: $P_o=-45$ dBm
f=1.0 GHz: $P_o=-55$ dBm
f=1.8 GHz: $P_o=-70$ dBm

Note: 0 dBm=1 mW in 50 Ω.
Construction of the calibration generator is straightforward even for those with limited experience in building RF circuits. It is essential that close-tolerance (2.5 or 5%) polystyrene capacitors be used in positions C_1, C_2 and C_4. Inductor L_1 is wound as 3 turns 0.2 or 0.3 mm dia. enamelled copper wire through a small (3 to 5 mm long) ferrite bead. Be careful to avoid short-circuits between the windings as the enamel coating may be damaged when the wire is pulled through the hole in the bead.
The calibration generator is powered from a 6 V battery pack so that it can be used as a portable test instrument. Current consumption is less than 20 mA.

B

Parts list

Resistors (±5%):
R_1 = 2K2
R_2 = 33K
R_3 = 47K
R_4 = 1K0 R_5;R_6;R_7 = 22K
R_8 = 56R

Capacitors:
C_1 = 470p*
C_2 = 22p*
C_3 = 40p foil trimmer
C_4 = 1n0*
C_5 = 180p
C_6 = 22n
C_7 = 4p7
C_8 = 390p
C_9 = 4µ7; 16 V
C_{10} = 10n ceramic
* Polystyrene (Siemens Styroflex); tolerance ≦5%.

Inductor:
L_1 = see text.

Semiconductors:
D_1 = 1N4148
IC_1 = 74HCT74
T_1 = BF494
T_2 = BF981 or BF982
T_3 = BFG65 (Philips/Mullard)
T_4 = BC550B

Miscellaneous:
X_1 = 2 MHz quartz crystal; 30pF parallel resonance.
PCB Type 884054 (see Readers Services page).

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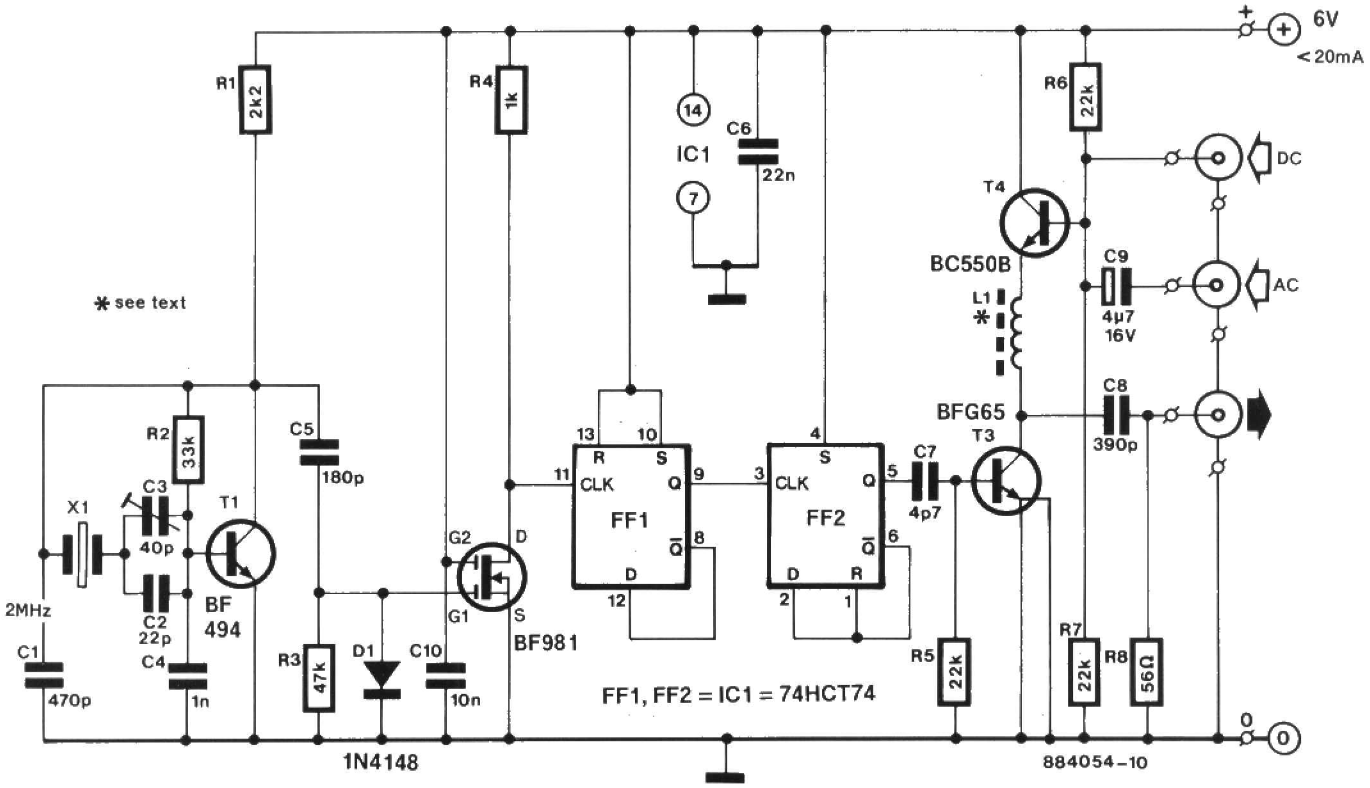

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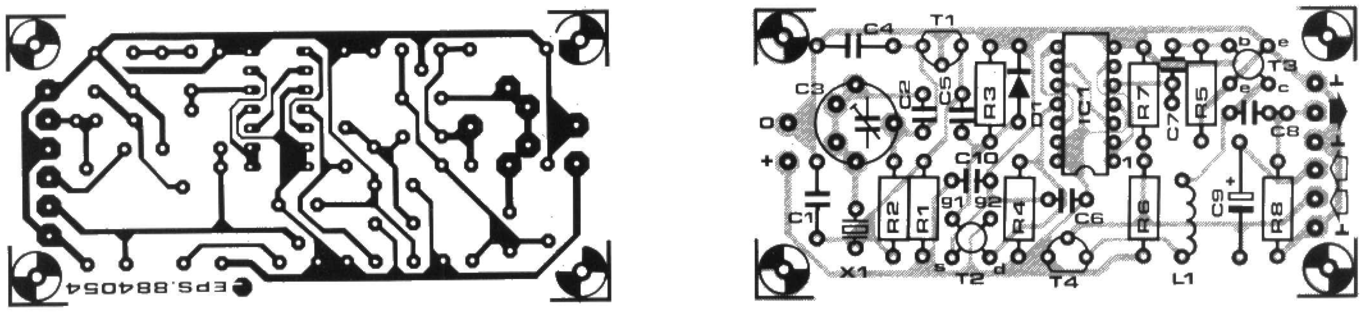

026 SIMPLE PHONO PREAMPLIFIER

This circuit shows that a preamplifier for magneto-dynamic cartridges can be relatively simple without seriously compromising compliance to the IEC standard in respect of frequency response. Compared to the RIAA standard, the IEC frequency curve has an additional roll-off point at 20 Hz—see Fig. 1.

The circuit diagram of Fig. 3 shows that input and output of the preamplifier based around the Type TL071 operational amplifier are direct coupled, making it possible to accurately define the previously mentioned roll-off by means of network R_2-C_3. Output offset of the preamplifier is about 3 mV. Output capacitor C_4 can be fitted if this offset voltage can not be handled by the input of the line or power amplifier.

For optimum compliance with the IEC frequency curve it is recommended to use close tolerance polystyrene (Siemens *Styroflex*) capacitors in positions C_1 and C_2, and an MKT capacitor in position C_3. Resistors are preferably high-stability metal film types from the E48 or E96 series, although less expensive and commonly available types from the E12 series may also be used with reasonable results when selected for the required resistance with the aid of a digital ohmmeter. It was with this in mind that R_2 has been dimensioned at 5K62 (E12: 5K6). This value gives a roll-off at 18.9 Hz instead of the required 20.0 Hz, so that the low-frequency response (up to 50 Hz) of the preamplifier deviates slightly from the IEC curve. The deviation, Δ, of the amplification with respect to the values set by the IEC is shown as a function of frequency in Fig. 2.

A prototype of the preamplifier built with the component values given in the circuit diagram gave the following test

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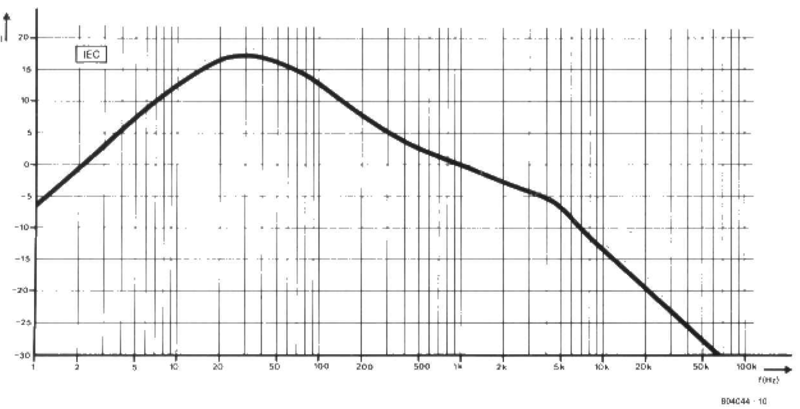

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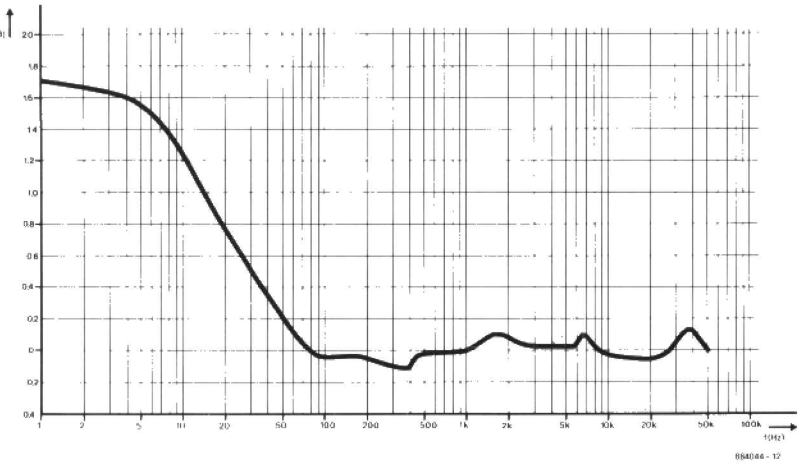

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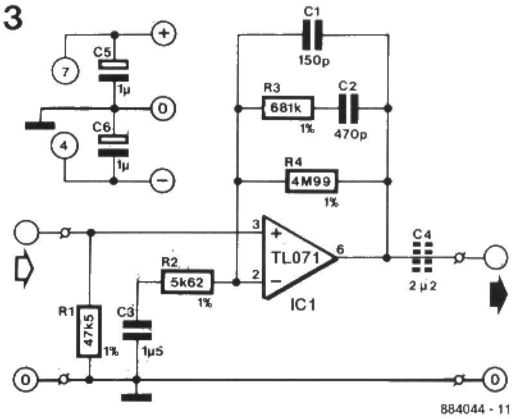

results: voltage gain 39 dB at 1 kHz; signal-to-noise ratio greater than 70 dB at 1 kHz and 100 mV output signal (up to 80 Hz: greater than 60 dB). The input was connected to a test generator which supplied 1 mV$_{rms}$ at an output impedance of 1 kΩ.

The circuit should be fed from a well-regulated symmetrical supply (preferably ±15 V, but ±12 V or ±8 V should also work). A suitable supply is simple to build around two integrated regulators such as the 78Lxx and 79Lxx types, which can step down supply voltages already available in the line or power amplifier. Current consumption of the preamplifier is only 2 mA.

027 ALTERNATING CURRENT SOURCE

One of the less known properties of field effect transistors is that some of these are electrically symmetrical, which means that the drain and source may be interchanged under certain conditions. This circuit is based on this phenonemenon, and feeds a constant alternating current through P2 when connected to an alternating voltage source.

The operation is best explained with reference to the curves of Fig. 2, and by assuming that a sinusoidal voltage is applied to terminals A and B.

When the drain of T1 is negative with respect to the source, D1 blocks, and forms a resistance that is considerably higher than that of R1. This has virtually no voltage on it, so that V_{GS} = 0 V. This means that I_D is constant at about 19 mA when V_{DS}>8 V (see Fig. 2a). It should be noted that the curves and values of I_D and V_{DS} are typical, and may deviate depending on the FET used (A, B or C suffix). When the drain of T1 is positive with respect to the source, D1 conducts. Provided P1 is adjusted such that the voltage on it equals V_D, there is, again, no voltage difference between the gate and the source, so that the FET functions as a current source as shown above.

The constant alternating current supplied by the circuit can be defined by fitting small resistors in the drain and source lines, so that V_{GS} is set to values other than 0 V. The input voltage range of the current source is 6 V$_{rms}$ to 18 V$_{rms}$.

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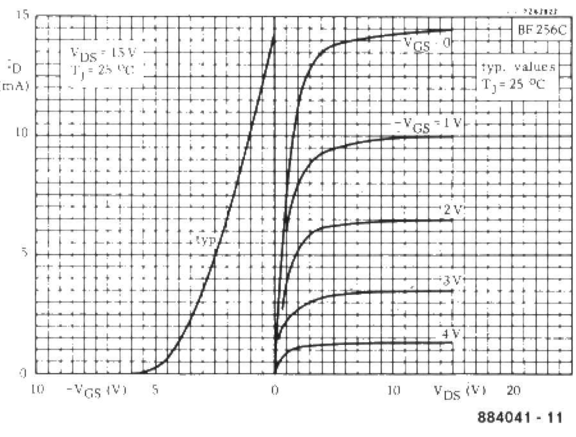

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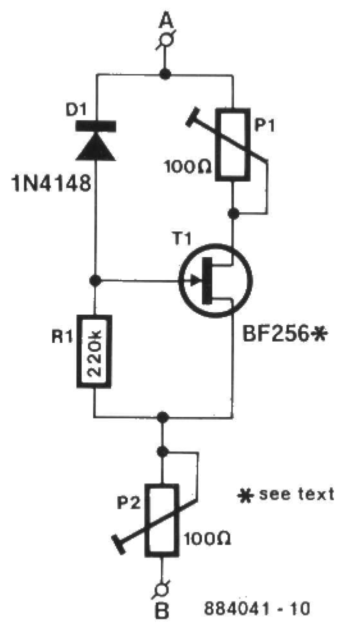
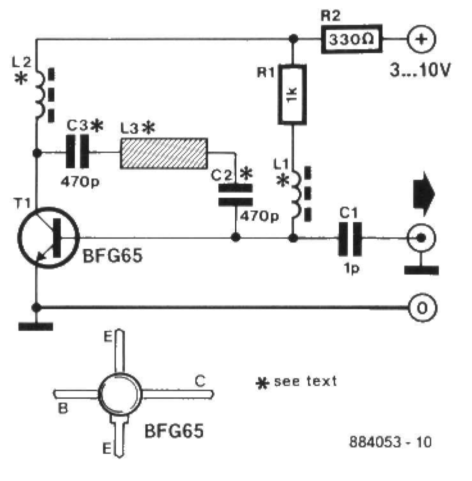
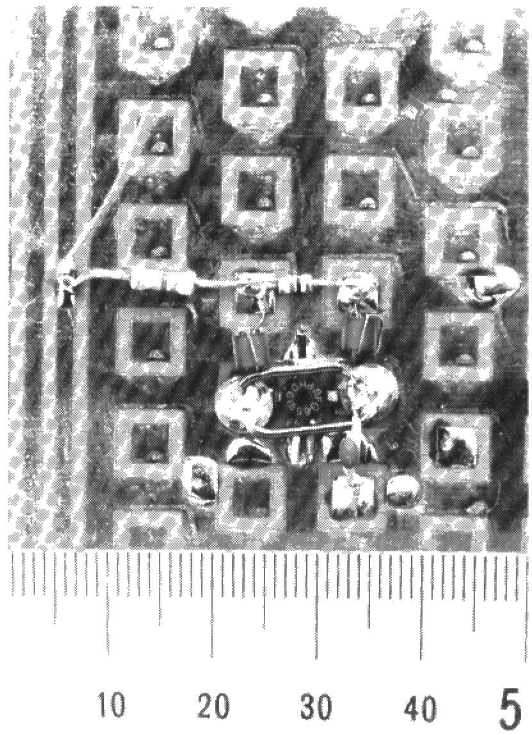


028 VOLTAGE-CONTROLLED SHF OSCILLATOR

This oscillator supplies an output level between −10 dBm and +3 dBm, and can be tuned between 1250 MHz and 1800 MHz simply by varying the supply voltage. Operation of the circuit is based on the fact that the transition frequency, f_T, of the BFG65 is reduced when the collector current rises above 10 mA. The oscillation frequency is also determined by the physical layout of inductor L_3, which is a strip line made from two parallel running lengths of 1 mm dia. silver plated wire. The length is established experimentally, starting from 13 mm. Chokes L_1 and L_2 are 3 turns of thin enamelled copper wire (dia. 0.2 or 0.3 mm) through a small (3 mm) ferrite bead. Capacitors C_2 and C_3 are leadless ceramic types (rectangular or disc).

The SHF test oscillator is ideal for quickly finding the maximum usable input frequency of, for instance, a frequency meter specified to reach up to 1.2 GHz. In addition, it can be used for testing RF input sections in indoor units for satellite TV reception.





029 COMPUTER-DRIVEN POWER CONTROLLER

This circuit enables a computer to control the power supplied to a mains operated device (lamp, heater, drill, etc.) in 256 steps. Variation of power is achieved by controlling the voltage supplied to the load (R_L in the circuit diagram of Fig. 2). A conventional power regulator is used here, composed of a triac and a simple associated circuit to control the phase angle at which the triac is triggered.

The power supply and mains trigger circuitry are shown in Fig. 1. The circuit around T_1. . .T_4 incl. and IC_1 is a zero-crossing detector which produces an active high pulse every time the mains voltage is zero. Opto-coupler IC_1 insulates the rest of the circuit from the mains.

With reference to Fig. 2, Schmitt-trigger N_1 inverts the zero-crossing pulses, causing 8-bit binary down counter IC_2 to load the 8-bit word applied to counter preset (jam) inputs JØ. . .J7. The counter is decremented one count by each clock pulse supplied by oscillator N_2. When counter state nought is reached, output $\overline{ZD}$ goes low, and N_3 inhibits further clocking of IC_2. Simultaneously, N_4 produces an output pulse, so that T_5 conducts and fires the triac.

As the triac is only fired when IC_2 counts to zero, the instant at which this happens depends on the value of the 8-

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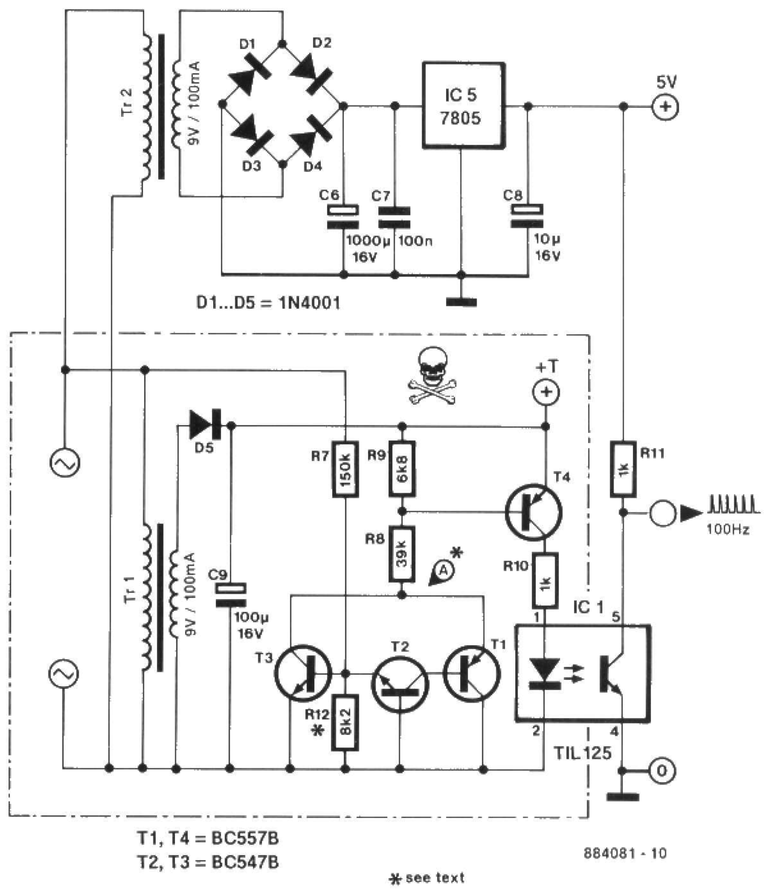

bit control word received from a computer. Hence, the time that lapses between the zero crossing instant and the triac firing instant is a function of the magnitude of the control word. The greater the 8-bit word, the greater the phase angle, and the less power is delivered to the load.
Inductor L_1 suppresses RF interference caused by the triac, and should be able to carry at least 5 A. The triac in this circuit can be a TIC206D (4 A) or a TIC216D (5 A). Other types may be used if these are known to trigger at a gate current of less than 10 mA. The value of R_{12} is determined empirically, and should be as high as possible without causing the disappearance, on point A, of pulses with an amplitude of 5 V_p.
The only adjustment required is that of P_1. If complete switching off of the load is required, this preset is adjusted for 0 V indicated by an AC voltmeter connected instead of the load, with data FF_H (255_{10}) written to the power controller. If regulation from 0 V onwards is not desired, P_1 is adjusted so that the meter reads the required minimum voltage. When writing programmes for the power controller, it should be remembered that the power delivered to the load is an inverse function of the value written into the computer's output port.

Safety precautions:
The shaded parts in the circuit diagrams are operated at mains potential, and must never be touched while the unit is being powered. Great attention should be paid to proper insulation in the selecting and mounting of the parts within the shaded areas. It is strongly recommended to bend the pins of the optocoupler away from the package to ensure an insulation distance of at least 6 mm.

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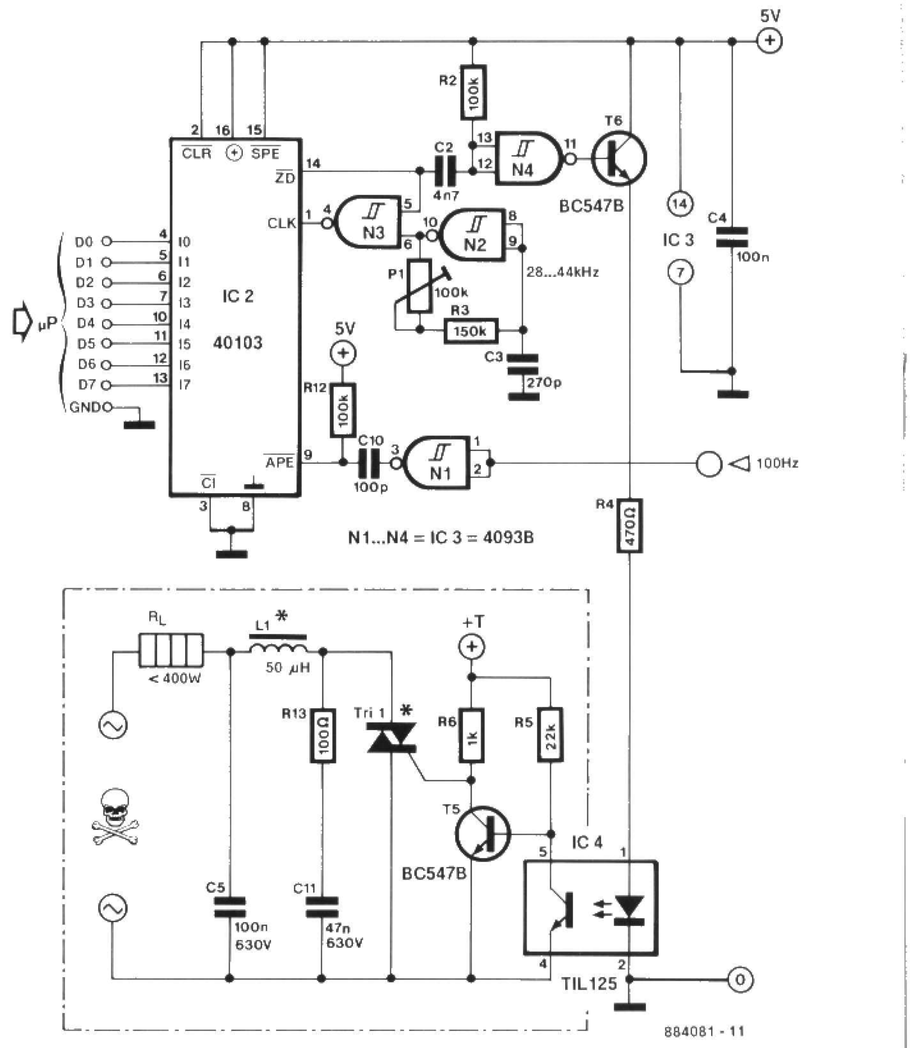
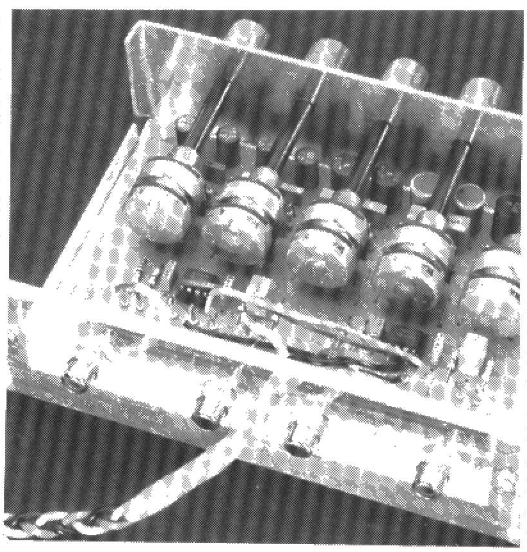


Finally, it should be noted that the circuit may not operate correctly with loads below about 40 W, and that writing 00_H to the data input has the same result as FF_H, namely minimum voltage applied to the load. Regulation efectively starts with data 01_H.

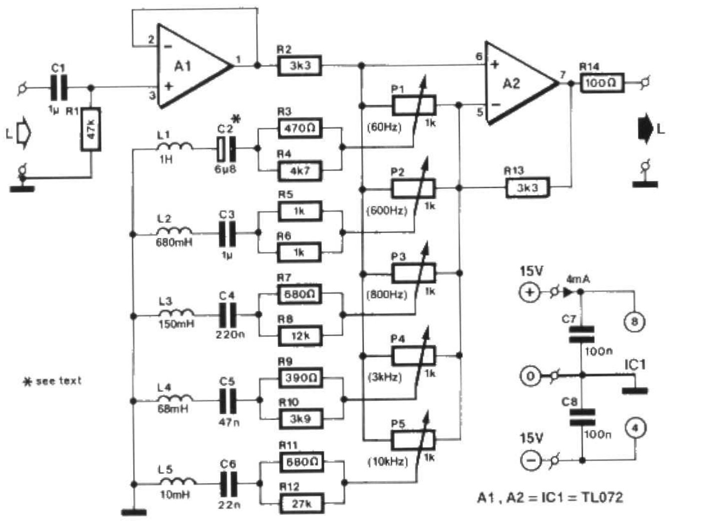
030

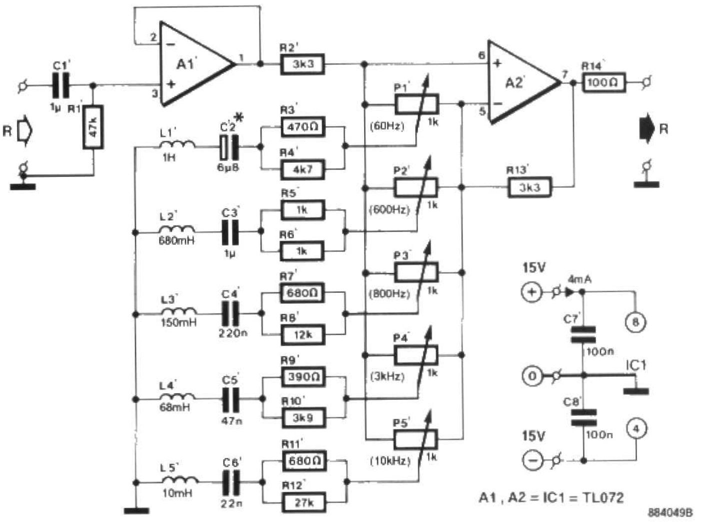
FIVE-BAND STEREO GRAPHIC EQUALIZER

This design of a stereo equalizer is fairly unusual because it is based on inductive feedback. In theory, the feedback circuit around opamp A_1 would provide 15 dB amplification or attenuation of each frequency range, but in practice only about 13 dB is attainable owing to losses in the inductors. A virtually flat frequency response is obtained when all five potentiometers P_1 to P_5 are set to the centre position (0 dB). Total control range of the unit is about 33 dB.
The TL072 dual opamp in each channel is a trade-off between cost and performance in respect of noise and distortion. Set to 0 dB gain, a prototype of the equalizer produced 0.04% distortion at an input signal of 1 kHz; 1 V, and 0.13% at 5 and 10 kHz. Distortion is highest when the test frequency lies within one band that is fully attenuated while the



other four are set to maximum gain. In this condition, test measurements resulted in a maximum distortion of 1.5%, which is certainly tolerable given the simplicity of the circuit. Signal-to-noise ratio is greater than 90 dB at an input amplitude of 1 V.
The frequency response curves were obtained with the following settings:
curve 1: all controls set to maximum;
curve 2: 4 controls set to 0 dB, and 1 to maximum;
curve 3: 4 controls set to 0 dB, and 1 to minimum;
curve 4: all controls set to minimum.
Due attention should be paid to the DC resistance of the inductors. The total resistance of the inductor and series resistors in each feedback network should remain 680 Ω, so that R_3 to R_{12} incl. may have to be dimensioned dif-





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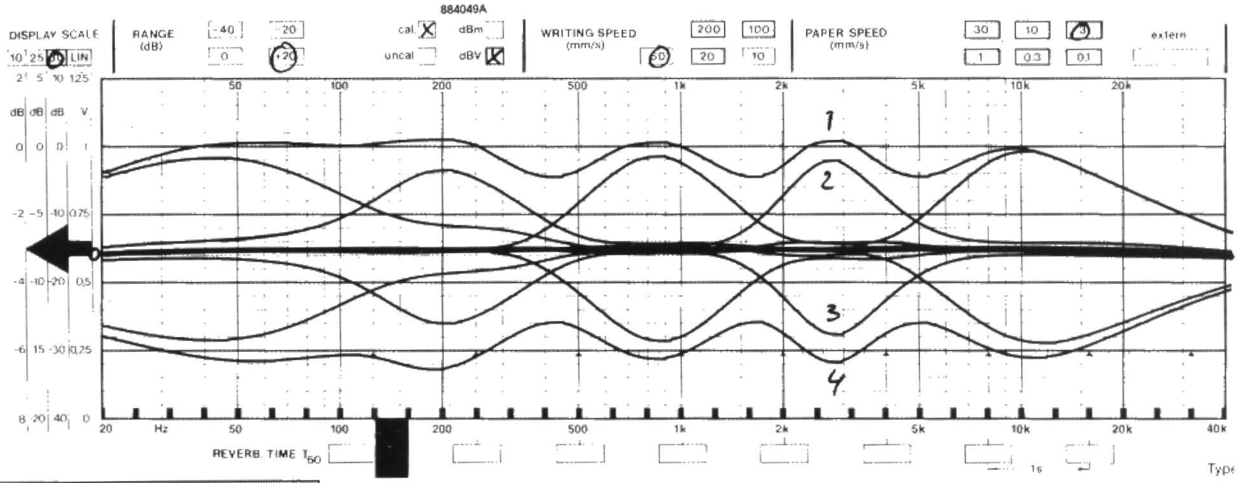


Parts list

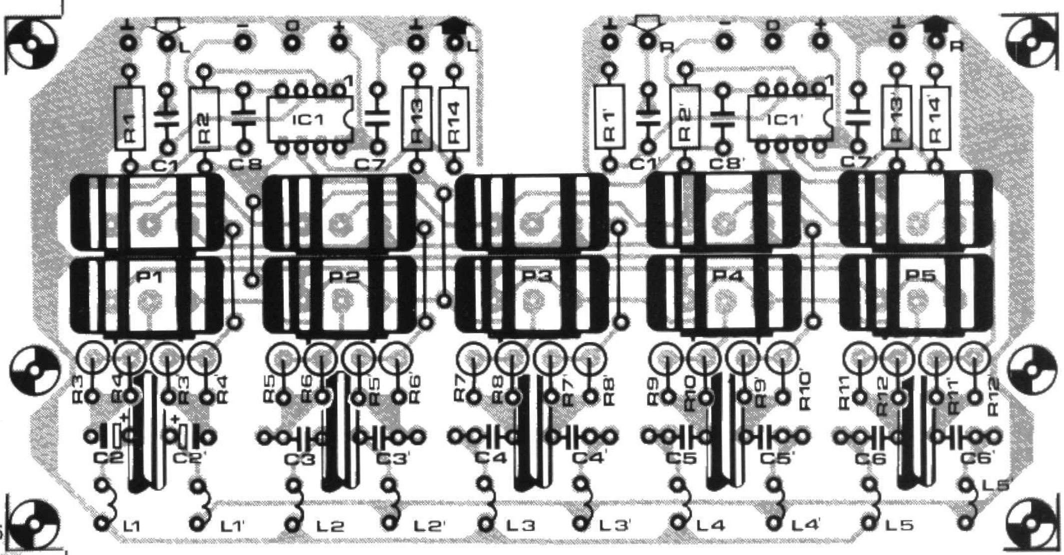
Resistors (±5%):
R1;R1' = 47K
R2;R2';R13;R13' = 3K3
R14;R14' = 100R
P1 . . . P5 incl. = 1K0 stereo linear potentiometer for PCB mounting.
• The following values are given as guidance only (see text):
R3;R3' = 470R
R4;R4' = 4K7
R5;R5';R6;R6' = 1K0
R7;R7';R11;R11' = 680R
R8;R8' = 12K
R9;R9' = 390R
R10;R10' = 3K9
R12;R12' = 27K

Capacitors:
C1;C1';C3;C3' = 1μ0 MKT
C2;C2' = 6μ8; bead tantalum
C4;C4' = 220n
C5;C5' = 47n
C6;C6' = 22n
C7;C7';C8;C8' = 100n

Inductors:
L1;L1' = 1 H, e.g. Toko Type 293LY-105 (Cirkit stock no. 34-10513).
L2;L2' = 680 mH, e.g. Toko Type 293LY-684 (Cirkit stock no. 34-68413).
L3;L3' = 150 mH, e.g. Toko Type 293LY-154 (Cirkit stock no. 34-15413).
L4;L4' = 68 mH, e.g. Toko Type 181LY-683 (Cirkit stock no. 34-68302).
L5;L5' = 10 mH, e.g. Toko Type 181LY-103 (Cirkit stock no. 34-10302).

Semiconductors:
IC1;IC1' = TL072

Miscellaneous:
PCB Type 884049 (see Readers Services page).

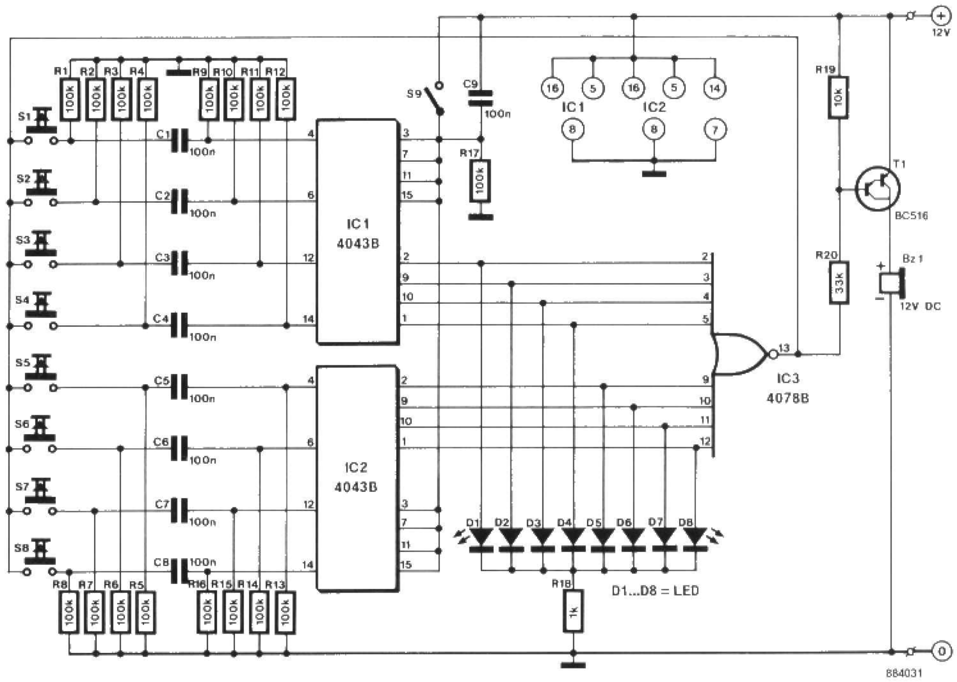


ferently than shown in the circuit diagram. Always measure the resistance of the inductors used, and then calculate the value of the resistor required to obtain a total of 680 Ω. Example: a Type 239LY-154 150 mH inductor from Toko was found to have a DC resistance of 37 Ω, requiring a series resistor of 680 − 37 = 643 Ω. This value is approximated with the aid of a 680 Ω and 12 kΩ resistor in parallel (R7-R8 in the circuit diagram). Ferrite-encapsulated inductors are recommended to reduce magnetic coupling, and to keep crosstalk at relatively high frequencies down to an acceptable level (< −60 dB at 10 kHz).

031 QUIZ TIMER

Here is a simple 'who's the first' circuit that can be used in quiz games with up to eight participants or groups of participants. The circuit indicates the first one to press his key by a glowing LED against his number or any other identification used in the quiz or game. At the same time, the circuit gives an audible indication that some key has been pressed. The RESET key enables the quiz master to restore the circuit's original state before updating the score and proceeding with the next question or assignment.

After RESET has been pressed, the eight R-S bistables in IC_1 and IC_2 are reset. The Q outputs all go logic low and, consequently, the output of IC_3 goes logic high. The circuit is now ready to be operated. For example, if S_1 is pressed first, the first bistable is set and output Q1 goes high. The output of IC_3 pulls the common line of keys $S_1 \ldots S_8$ incl. logic low to prevent more bistables being set. Hence, Q1 is the only output that is logic high. This condition is indicated by LED D_1. Simultaneously, T_1 is biased and switches on the buzzer to attract the attention of the quiz master. Capacitors C_1 to C_8 incl. prevent the bistables being set permanently if a key is kept pressed for a long time. Finally, S_9 is pressed to reset the circuit. This causes all Q outputs to be made logic low, and the common key line high, returning the circuit to its original condition.



The circuit is not critical in respect of supply voltage, which is preferably the working voltage of the active piezo-buzzer (6 V or 12 V). Current consumption in the de-activated state is less than 1 mA, while less than 25 mA is drawn when one of the LEDs is illuminated. The supply voltage need not be regulated, making it possible to use an inexpensive mains adapter of the DC type.

032 HEADLIGHTS INDICATOR

It is never advisable to leave a car's headlights on for long periods when the engine is not running. Yet, especially during the winter months, many of us inadvertently do this. The indicator described here helps to prevent you suffering the consequent and inevitable flat battery.

In its simplest form, the indicator consists of a d.c. buzzer and a diode as shown in Fig. 1a. With the headlights on (S_3 closed), the interior light (L_1) does not come on until one of the front doors is opened (when either S_1 or S_2 closes). At the same time, the buzzer is energized and sounds. Either closing the door or switching off the headlights (S_3) turns the buzzer off.

The diode in series with the buzzer is necessary because B is normally at +12 V via L_1 and A at ground via the headlights (L_2—only one shown here).

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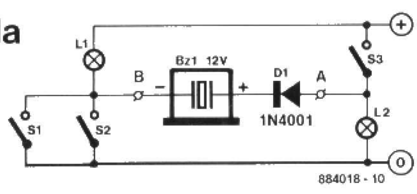

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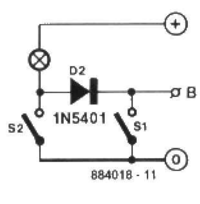


The buzzer should then not sound, of course.

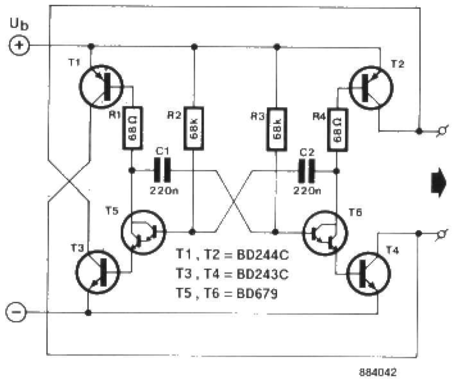
A slight modification to the circuit, enabling it to operate only on the switch in the driver's door, is shown in Fig. 1b (where it is assumed that S_1 is the relevant switch).

The buzzer draws a current of only about 10 mA when energized.

033 POWER MULTIVIBRATOR

This simple multivibrator circuit is remarkable for its high efficiency and ability to drive relatively heavy loads. The circuit supplies a symmetrical rectangular signal that floats with respect to the supply voltage. An astable multivibrator is formed by T_5, T_6, R_1, R_2, C_1 and C_2. The collector currents of T_5 and T_6 drive T_1 and T_2 respectively, while the emitter currents drive T_3 and T_4 respectively. Current limiting may be dimensioned to requirement by changing R_1. It should be noted, however, that the transistors may carry relatively high currents. Their current amplification, h_{FE}, is, therefore, fairly low, so that the current limit point can be approximated with $h_{FE(max)}(U_b-1.4)/R_1$. With $R_1=68\ \Omega$ as shown in the circuit diagram, the multivibrator can be used for switching loads up to about 3 A.

Output frequency of the oscillator is approximated by $0.7/(R_2C)$, and is about 53 Hz with $R_2=68\ k\Omega$, $C=C_1=C_2=220$ nF and $U_b=12$ V (14 V: 50 Hz). One of the many applications of



the power multivibrator is a battery-operated mains converter. Its outputs are then connected to the low-voltage secondary winding of a mains transformer. A prototype of the multivibrator was dimensioned for relatively high output current at 50 Hz by fitting $R_1=33\ \Omega$; $R_2=2\times68\ k\Omega$ in parallel, and $C=2\times220$ nF in parallel. Connected to a 9.5 V; 5 A mains transformer, it powered a 40 W mains bulb with a rectangular voltage of nearly 240 V_{rms}. Supply voltage and current consumption were 14 V and 6 A respectively, yielding an acceptable efficiency of about 40%. Quiescent current consumption of the circuit is determined by R_1, and was 0.3 A in the test set-up.

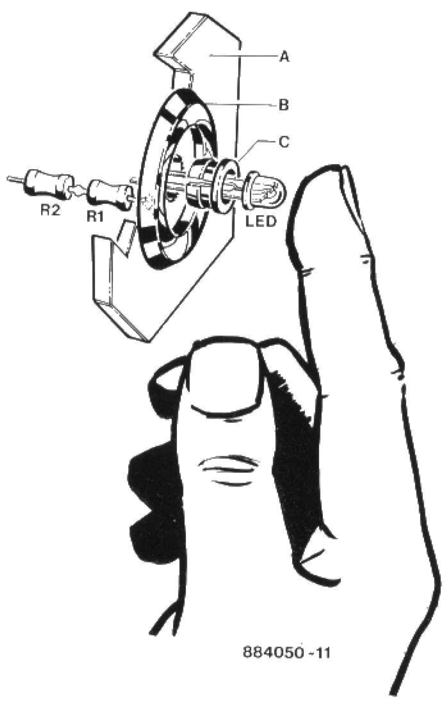
When the multivibrator is used for driving an inductive load, as in the above application, each output transistor must be protected from inductive voltage peaks by two fast high-current diodes fitted in reverse across the collector and emitter terminals.

034 TOUCH SENSITIVE LIGHT SWITCH

This low-cost circuit enables turning room lights on and off simply by touching a round metal sensor. The light is turned on by briefly touching the sensor, and off again by touching it slightly longer. With reference to the circuit diagram, when the sensor is briefly touched, hum and noise induced on the body are amplified by cascaded gates N_1, N_2 and N_3. A pulse train with a swing of nearly the supply voltage (4.7 V) and a frequency equal to that of the mains voltage (50 or 60 Hz) is applied to a bistable set up around N_4 and N_5. C_2 is charged via D_2, and the bistable latches in a high output state. Triac Tri_1 is triggered via driver T_1, so that the lamp lights.

When the sensor is touched for about 2 seconds or longer, the pulse train charges C_1 via R_5 and D_1. Inverter N_6 pulls the input of N_4 low when the voltage on C_1 is sufficiently high. Bistable N_4-N_5 toggles and T_1 breaks the gate current for the triac, so that the lamp is turned off. The circuit also works in a relatively noise-free environment. When the user forms a relatively low resistance to ground, the input of N_1 is effectively pulled low by R_1-R_2, whose total resistance is low relative to

1



R_3-R_4. The effect of this on the bistable and triac circuit is similar to that outlined above.

A suggested construction of the sensor and LED is shown in the accompanying drawing. The LED is fitted in a plastic holder, and in the dark indicates the location of the light switch. The LED holder (C) is secured in the side or top panel (A) of the ABS enclosure that houses the light switch circuit. The LED is located in a thin aluminium or brass washer (B), which is connected to R_1, and glued onto the outside surface of the plate. In the interest of safety, it is recommended to observe a minimum distance of 7 mm between the LED and R_1. In this context, constructors are urgently advised **not** to use a metal or metallized LED holder as the sensor. Also, **never** replace R_1 and R_2 with a single 4M7 resistor.

Since this circuit is connected direct to the mains, it must be fitted in a safe and sound ABS enclosure that is impossible to open without it being deliberately damaged. Once more we advise that the presence of the mains voltage is a serious source of danger, so that the first and foremost concern of every constructor should be absolute safety.

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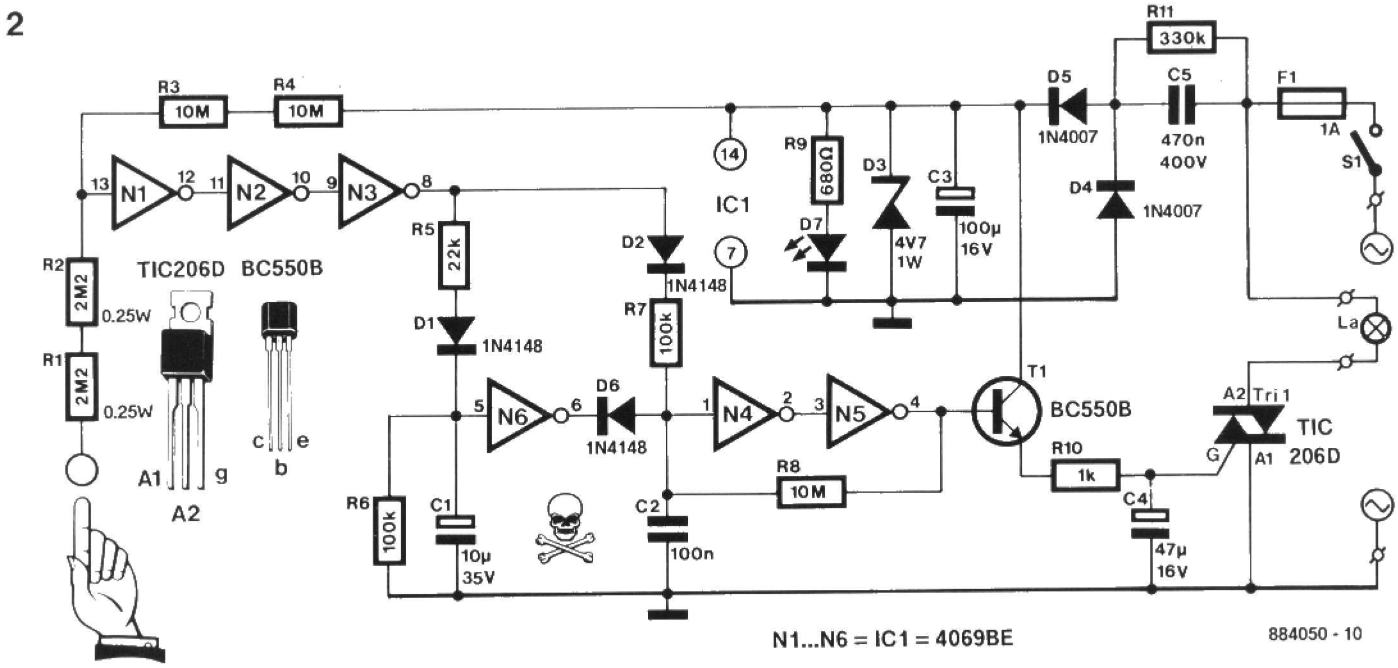


035 PRINTER SHARING BOX

This simple circuit makes it possible to connect two computers to a single printer. Toggle switch S_2 selects the relevant computer by applying the appropriate logic level to the enable inputs, $\overline{G}$, of octal bus transceivers Type 74LS641 ($IC_1 \ldots IC_4$ incl.). The direction input, DIR, of these is hard-wired to +5 V, so that the data direction is from *An* to B*n*. When $\overline{G}$ is logic high, the buffers are switched to the high impedance state, so that chip outputs can be connected to form a bus structure. With this in mind it is relatively simple to see that the circuit is the electronic equivalent of a 16-way toggle switch.

Input BUSY of the non-used computer is held logic high to prevent this machine attempting to send data when the other computer is accessing the printer. The 74LS641 was chosen because it has open-collector outputs — the reason for this should be clear when it is remembered that the Centronics standard dictates the presence of pull-up resistors in the printer. The 74LS641s, of course, need pull-up resistors at the computer side also, and these are formed by resistor networks R_3, R_4, R_6 and R_7.

A RESET switch, S_1, is provided to clear the printer buffer by means of an $\overline{\text{INPUT-PRIME}}$ pulse should the user find out that the wrong file is being printed. This reset option is definitely neater than switching off the printer completely to correct the error.

The circuit is conveniently powered from the 5 V supply in the printer. In most cases, this supply voltage is available on pin 18 of the 36-way Centronics input connector, but this would have to be ascertained by measuring and reference to the printer manual. It is recommended to connect +5 V to non-used pins 15 and 34 also to distribute the current over several wires in the Centronics cable. Once again, check the

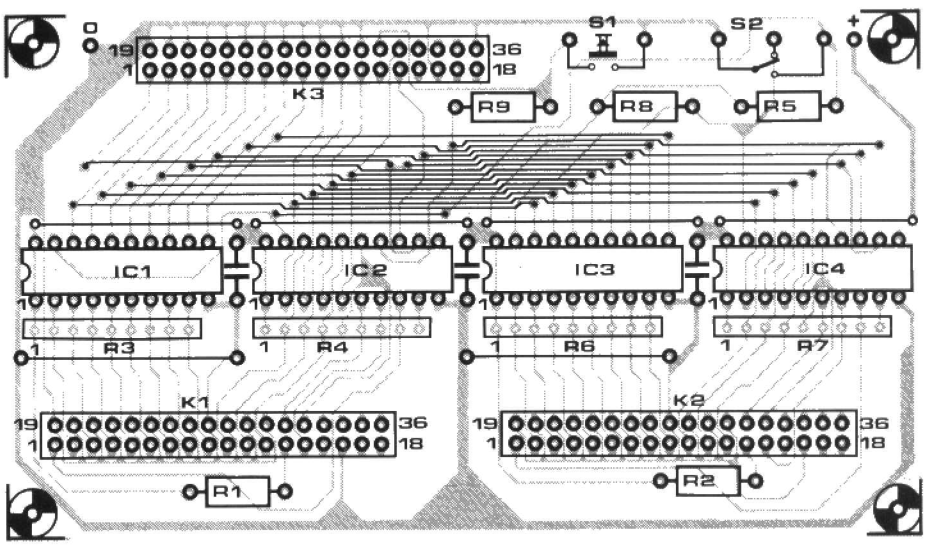
Parts list

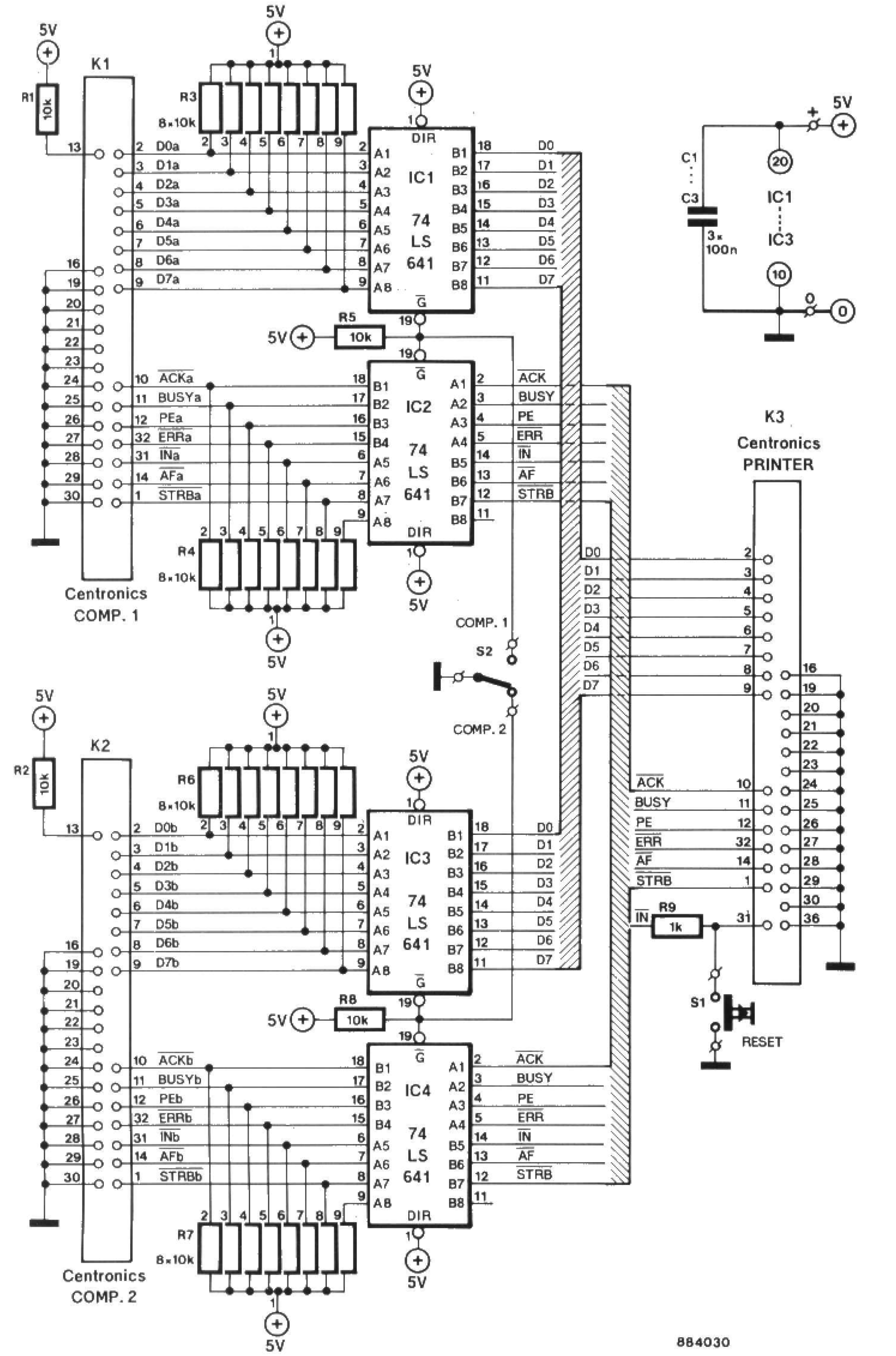
Resistors (±5%):
R_1;R_2;R_5;R_8 = 10K
R_3;R_4;R_6;R_7 = 8-way SIL resistor network 10K
R_9 = 1K0 1

Capacitors:
C_1;C_2;C_3 = 22n

Semiconductors:
$IC_1 \ldots IC_4$ incl. = 74LS641

Miscellaneous:
S_1 = miniature toggle switch.
S_2 = miniature toggle switch.
K_1;K_2;K_3 = 36-way terminal strip block.
PCB Type 884030 (not available ready-made through the Readers Services).





printer manual to see whether these pins are actually available for this purpose.

The printer sharing box will generally be located close to the printer. PCB connectors K_1, K_2 and K_3 are 36-way straight headers. Three cables are required for connecting the completed PCB between the computers and the printer.

Two 10—15 cm long adaptor cables are made from flatcable with IDC (press-on) connectors at either end. One end is terminated in a 36-way IDC socket for plugging onto the PCB header, the other in a female IDC Centronics socket (blue ribbon type) for receiving the printer cable.

The short printer output cable is composed of a female 36-way IDC socket as above, and a 36-way male Centronics plug.

Current consumption of the printer switch is about 200 mA.

036

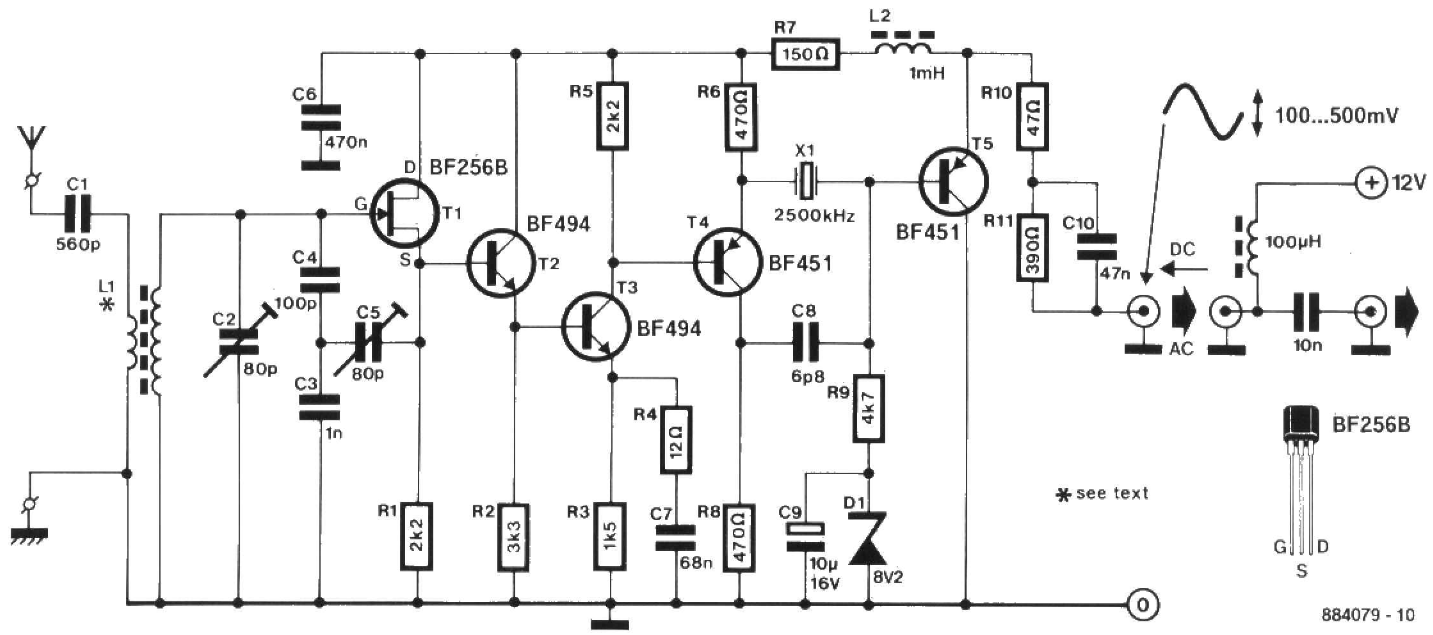
OMA-2500 TIME STANDARD RECEIVER

OMA-2500 is a 1 kW time standard transmitter on 2500 kHz. The station is located in Liblice, Czechoslovakia, and is operated by the Astronomical Institute of the Chechoslovak Academy of Sciences.

Contrary to time standard transmitters in the VLF band (DCF77, HBF), modulation is pure AM instead of a combination of AM and PSK or FSK. This means that the seconds pips transmitted by OMA-2500 are free from phase noise, which is a must for some types of PLL, particularly in communications equipment, where the 2500 kHz signal supplied by a time standard receiver is used for generating or deriving other frequencies of equal stability.

Transistor T_1 is configured as a regenerative buffer which acts as an active filter with an effective Q (quality) factor of about 1,000 at a 3 dB bandwidth of 2.5 kHz. The received signal is further raised in amplifier T_2-T_3 before it is applied to active crystal filter T_4-X_1 which ensures a 3 dB bandwidth of about 500 Hz. Output amplitude of the receiver is sufficient for driving almost any type of simple PLL. The receiver is powered via the downlead cable at the output to enable it to be mounted in a noise-free environment.

Inductor L_1 is wound as 2 turns (primary) and 50 turns (secondary) of 0.3 mm dia. enamelled copper wire on a Type T50-2 core. Quartz crystal X_1 is a



2500 kHz type for series resonance. Construction of the receiver should follow the standard rules for RF circuits: keep all connections as short as possible, and use ample screening and decoupling.

Adjustment: set a function generator to 2.5 MHz at $U_o = 10$ mV. Connect the output to C_1. Connect an AC-coupled oscilloscope to the source of T_1, and peak C_2. It may be necessary to reduce or increase the number of windings of the secondary of L_1 to obtain resonance at 2.5 MHz. Reduce the signal amplitude and redo the adjustment of the trimmer. Disconnect the function generator, and connect the aerial. Connect the scope to the output of the circuit. Peak C_5 for optimum amplitude of the AM signal, but make sure that this is not greater than 500 mV. Remember that T_1 is a regenerative stage, so that the settings of C_2 and C_5 interact. If necessary, readjust the trimmers to ensure that the signal at the collector of T_3 is stable, and not clipped during night-time reception, when OMA-2500 is received with high field strength throughout Europe. Daytime reception in western and northern Europe will mostly range from poor to just usable, depending on propagation conditions and location of the receiver.

The circuit is fed from 12 V, and consumes about 10 mA. Finally, bear in mind that a good aerial (long wire or rhombic quad) is imperative for reliable reception.

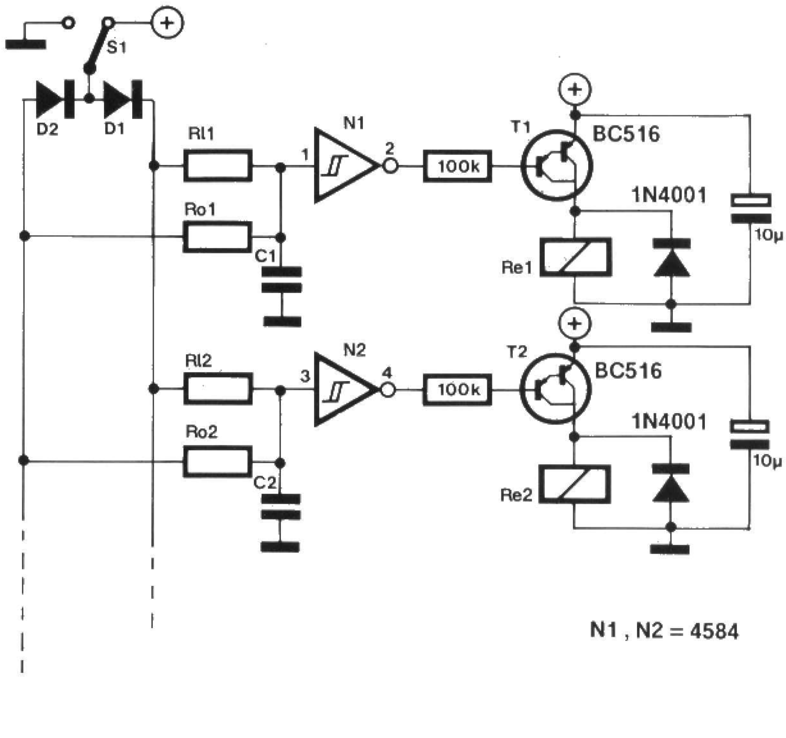
037 PROGRAMMABLE SWITCHING SEQUENCE

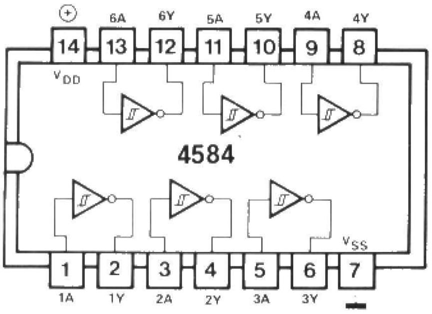
The proposed control circuit is shown in the diagram as containing two relays, but this number may be increased if necessary. The switching sequence is determined by the time delay of an RC network at the input of a gate that is used to energize a relay via a darlington transistor.

When S_1 is connected to the supply voltage (as shown in the diagram), the input capacitor, C_1, C_2, . . . , begins to charge via a resistor, Ri_1, Ri_2, . . . , and diode D_1. After a given time, depending on the time constant of the relevant RC combination, the voltage across the capacitor has reached a value sufficient to toggle the gate. The relevant transistor is then switched on, and the relay is energized.

By giving the input of each gate a different time constant, the sequence of switching is determined.

When S_1 is switched to ground, the opposite happens. Diode D_1 is reverse-biased and the capacitors, C_1, C_2, . . . ,

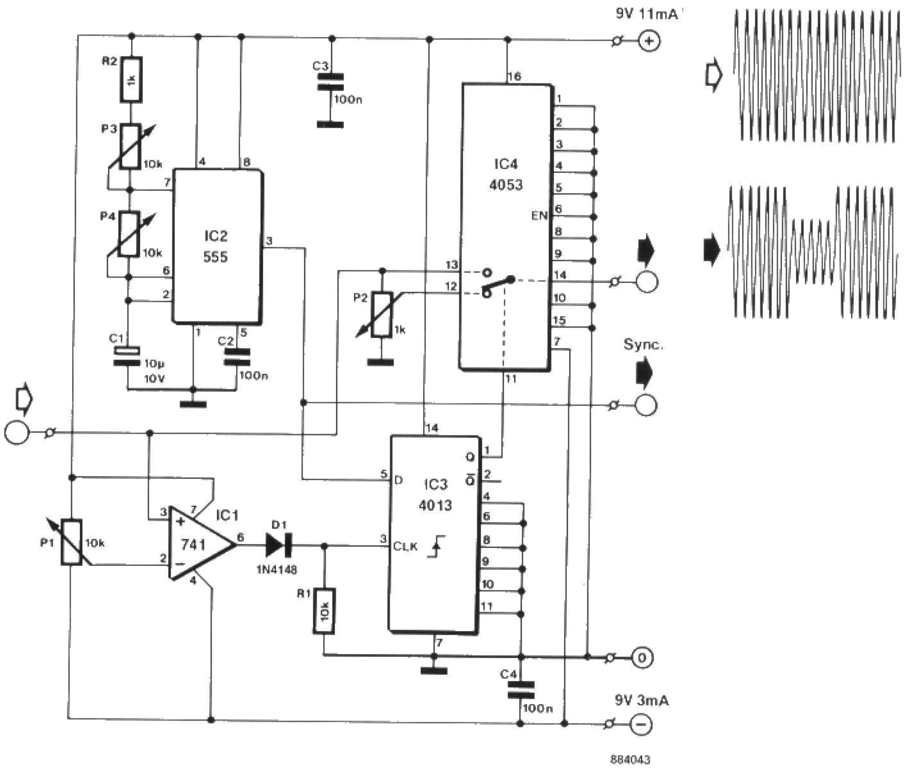




discharge via resistors R_{o1}, R_{o2}, . . . , and diode D_2. The discharge time constant determines how fast the capacitors can discharge and retoggle the gates. So, here again the switching sequence is determined by time constants. The gate with the shortest time constant will always toggle first.
The supply voltage may lie between 5 V and 15 V, but must, of course, be equal to the operating voltage of the relays. Furthermore, the BC516s must not switch more than 400 mA, and this again influences the choice of relay. A good, practical energizing current for the relays is 200 mA.
The values of resistors Ri and Ro may lie between 1 kΩ and 10 MΩ; the value of capacitors C_1, C_2, . . . , between 10 pF and 100 μF. Time constants exceeding 1,000 seconds create problems in practice, because the leakage current of the electrolytic capacitors then becomes comparable with the charging current. In general, choose the time constants so that two consecutive ones always differ by at least 0.1 s.

038 BURST GENERATOR

A burst generator is indispensable for testing the dynamic response of loudspeakers, and, in some cases, AF amplifiers. The fact that a number of cycles of a sinewave are applied to the loudspeaker under test, and not a continuous signal, eliminates the adverse effects of reverberation, reflection and echoes which are otherwise caused by the test room, and are almost inevitably picked up by the test microphone. In addition, the burst provides a good indication of the loudspeaker's performance in respect of voice coil transient response, resonance, and ringing.
The test signal provided by an external sinewave generator is switched on and off at or around the zero crossing, depending on the setting of phase control P_1. The pause amplitude can be set by P_2, while controls P_3 and P_4 are used for adjusting the duration of the pause and the burst respectively. It should be noted that the settings of these potentiometers interact, so that an oscilloscope is required for correct alignment. The duration of pause and burst is not related to the input signal. This means that the number of cycles supplied by the generator increases with the frequency of the sinewave applied to the input, unless, of course, P_3 and P_4 are re-adjusted.
Comparator IC_1 converts the sinewave at the input into a rectangular signal. The switching takes place at a specific instantaneous amplitude of the sinewave, set by P_1. The timing of the switching instant is arranged by astable mulitivibrator IC_2, and is copied in bistable IC_3 on the first positive edge of the sinewave, since this corresponds to the rising edge of the clock signal. Output Q goes high, so that the pole of electronic toggle switch IC_4 is connected to pin 12, and hence carries the attenuated sinewave burst.
The burst generator is not critical in respect of supply voltage, as long as this remains between ±5 V and ±9 V. Do not exceed ±9 V on penalty of damaging IC_4.



039 I/O EXTENSION FOR AMIGA 500

The Commodore Amiga is claimed to be a computer with plenty of facilities for extension circuits. The model 500, for instance, comes with no fewer than twelve connectors and sockets. There are, however, awkward constraints to the practical use of all these extension facilities. The SERIAL CONNECTOR is cumbersome to use with TTL circuits because of the ± 12 V logic levels on it. The use of the 86-PIN CONNECTOR on the machine is complex and risky because of the unbuffered connection to many internal signals. The one remaining option is the PARALLEL CONNECTOR, which can be extended to a maximum of 56 I/O lines as shown here, with the possibility to realize a bidirectional port.
The circuit was designed and built for the Amiga 500 computer. It is likely to work equally well on models 1000 and 2000, but this has not been tested in practice.

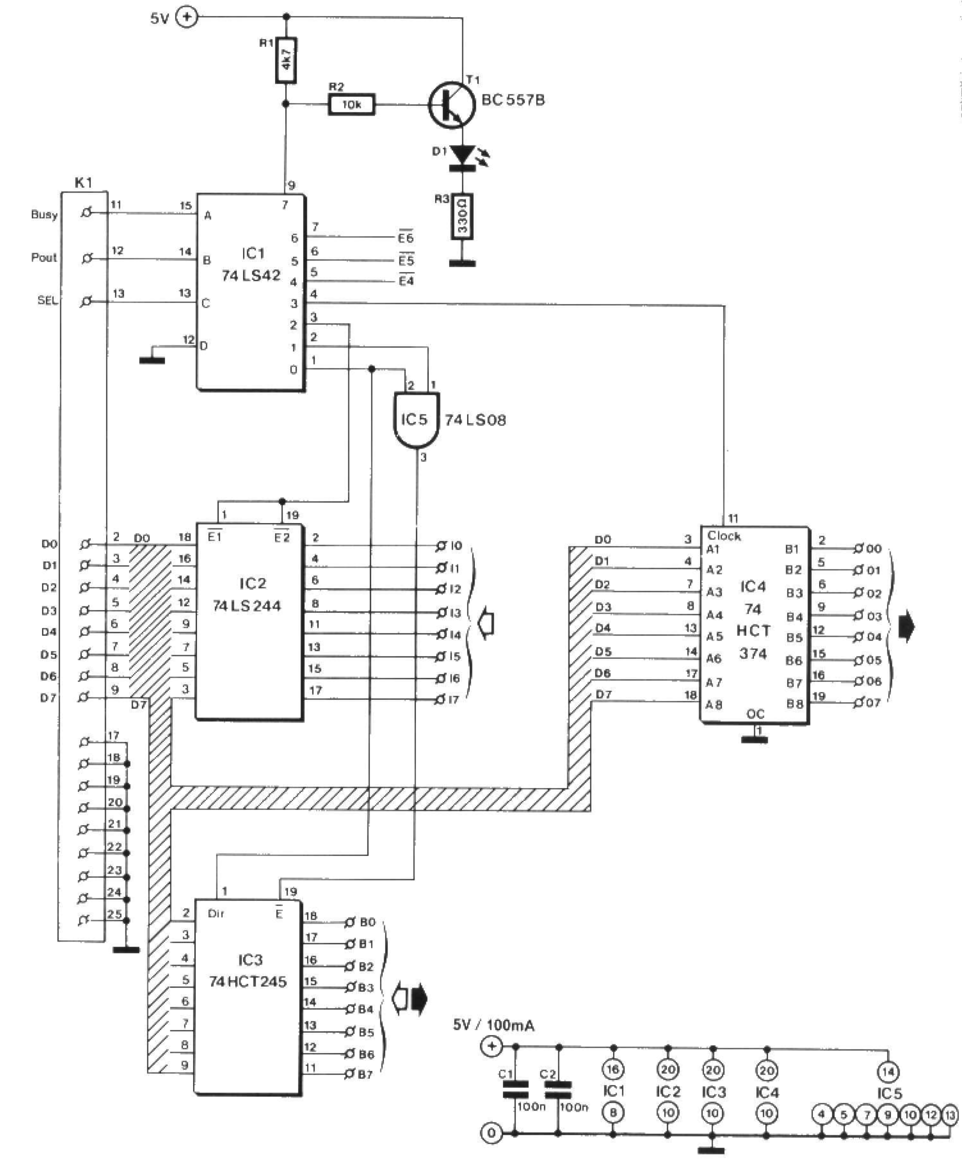
Output lines BUSY, P-OUT and SEL on the PARALLEL CONNECTOR can be programmed to supply a 3-bit address selection code which is applied to binary decoder IC$_1$. Octal bus buffer IC$_2$ is the input port at address 2, latch IC$_4$ the output port at address 3, and transceiver IC$_3$ the bidirectional port at addresses 0 (read) and 1 (write). The remaining 3 addresses (lines $\overline{E4}$, $\overline{E5}$ and $\overline{E6}$) can be used for $3\times 8=24$ additional I/O lines. Line 7 on IC$_1$ may not be used for selecting an input ot output port, and is used instead for driving READY LED D$_1$ when none of the ports on the I/O extension is being selected. It should be noted that IC$_3$ is not a latch, which means that it can only output data for as long as it is written to by the microprocessor. Output port IC$_4$ does have a latching function, so that datawords are kept stable on the outputs until overwritten by the microprocessor.
The accompanying listing is intended as a guide to writing software for the I/O extension. As an example of the practical use of the subroutines, instruction

a=1:n=123:GOSUB Wr <CR>

sends dataword 123_{10} to IC$_3$, which then functions as an output port. Conversely, instruction

a=2:GOSUB Rd:PRINT n <CR>

reads the dataword applied to IC$_3$, and prints it on screen.
Subroutine Init need only be called once at the beginning of the programming session. Input ports must not be written to. The I/O extension should be fed from a separate 5 V supply.



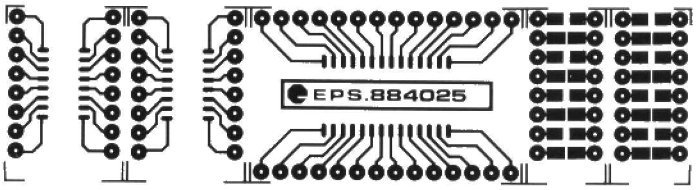

```
Init:                          'call once after power-on
  POKE 12571136&,199           'BUSY, P-OUT and SEL = output bits
  POKE 12570624&,255           'select address 7 (light READY LED)
  POKE 12575489&,0             'set port to input
RETURN


Rd:                            'load contents of address a in variable n
  POKE 12575489&,0             'set port to input
  POKE 12570624&,248+a         'select address a
  n=PEEK(12574977&)            'read value
  POKE 12570624&,255           'light READY LED
RETURN


Wr:                            'store variable n in address a
  POKE 12570624&,248+a         'select address a
  POKE 12575489&,255           'set port to output
  POKE 12574977&,n             'write value
  POKE 12570624&,255           'light READY LED
RETURN
```

040 UNIVERSAL SMD-TO-DIL ADAPTORS

An increasing number of electronic components, and in particular integrated circuits, is now only available as surface-mount devices (SMDs). Circuit design on the basis of these leadless, tiny, components invariably poses problems to many because there is no way to go round making a printed circuit board for building and testing prototypes. Making PCBs for SMD based designs is cumbersome and time-consuming. In many cases it will, therefore, be desirable to develop the circuit as it would have been done using ICs and components of standard size. The PCB adaptors introduced here make this possible. With the exception of the general-purpose type, they are slightly larger than ICs of normal size, but still fit in the generally adopted 0.1 in. raster.
The adaptor PCBs effectively enable a range of SMD ICs to be handled just as their normal-size equivalents, and so alleviate the plight of designing and etching a new PCB for every experiment or minor change to the circuit.
SMD ICs with 8, 14 or 16 pins are usually housed in a 'narrow' enclosure, and 16, 20, 24 and 28 pin types in a 'wide' enclosure.
The printed circuit board shown here allows making multiple adaptors that can be used for fitting:



Note: the printed circuit board shown here is available ready-made through the Readers Services under order no. 884025.

- Narrow SMD ICs with a maximum of 16 pins. For 8 and 14 pin types, the PCB can be cut to the required length.
- Wide SMD ICs with a maximum of 28 pins. PCB sections are cut off to the required length as above.
- SMA transistors, capacitors and resistors. These are arranged in a DIL configuration on a general-purpose adaptor to enable fitting networks and circuit sections as complete modules on a standard prototyping board. The size of this adaptor does not exceed that of a standard 16-pin integrated circuit.

Suitable lengths of terminal strip are pushed through the holes at the underside of the boards to create pins for fitting the modules in standard IC sockets.

TW

041 WIPER DELAY

This two-key wiper delay circuit is remarkable for its simplicity and ease of use. The wipe is started by pressing the SET switch, which also serves to adjust the length of the wipe interval. The circuit is turned off by pressing the RESET button.
The wiper delay shown in Fig. 1 consists of three opamps and a monostable multivibrator (MMV). Opamp A_1 is set up as a triangular wave generator, controlled by the output of the MMV. When this is low, a slowly rising sawtooth voltage appears at the output of A_1. The rise time of the sawtooth depends on R_2-C_3. Opamp A_3 compares the voltage across C_4 to the instantaneous sawtooth amplitude. The output of A_3 drops from 8 V to 0 V when the sawtooth voltage exceeds U_{C4}. This change in the output voltage of A_3 is delayed by R_6-C_6 and passed to A_2, so that the MMV is triggered somewhat later. The wipers are switched on via T_1 and Re when pin 3 on the 555 goes high. Also, C_3 is rapidly discharged via D_1 and R_1, while D_2 prevents the voltage across C_3 becoming positive. When the MMV output goes low, A_1 generates a new sawtooth period.

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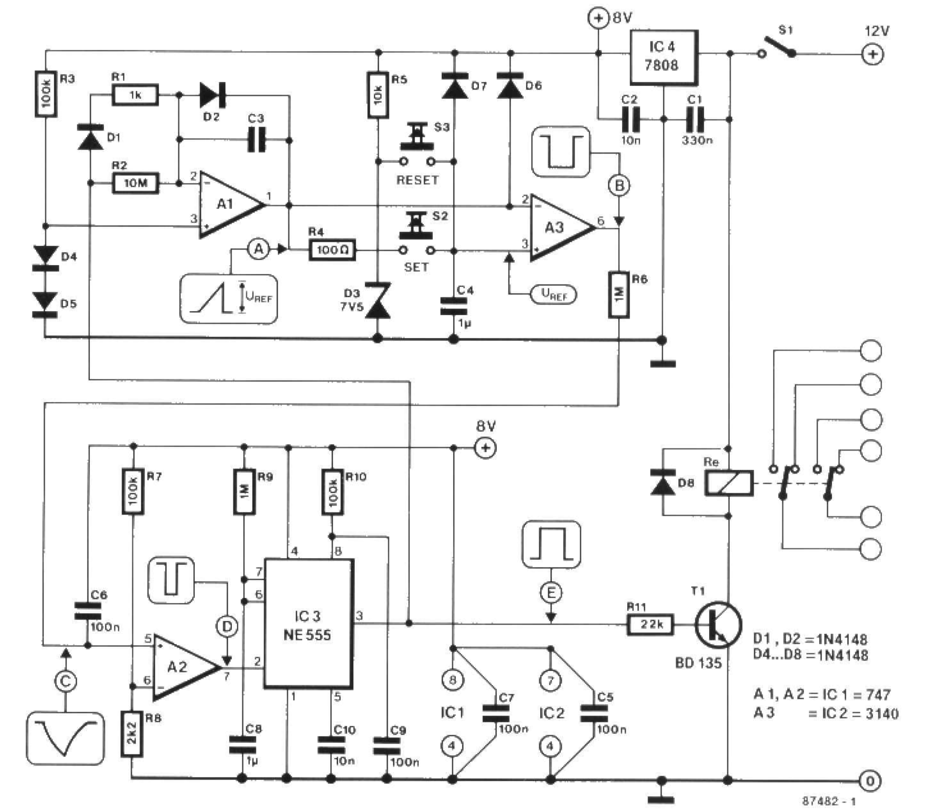

When the circuit is first switched on, C_4 is discharged, and the output of A_1 is slightly higher than 0 V due to $V_{[C\text{-}E]}$ of the internal output transistor. This causes the outputs of A_3 and A_2 to remain low, so that the wiper relay remains energized initially. When RESET is pressed, C_4 is charged via R_5, causing the the output of A_3 to go high, and the MMV to be stopped. The delay circuit around A_2 is necessary to prevent C_4 being discharged completely after pressing the SET button.
The relay contacts should be wired such that the dashboard switch is bypassed when the relay is energized, and that the hold switch, H, for the wiper motor is opened—see Fig. 2. Due attention should be given to the correct connection of the hold switch on penalty of short-circuiting the car battery.

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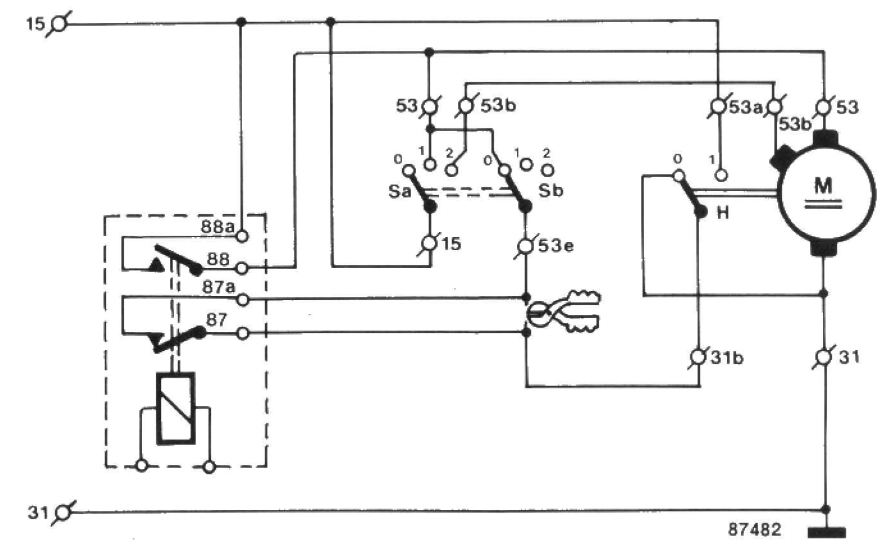
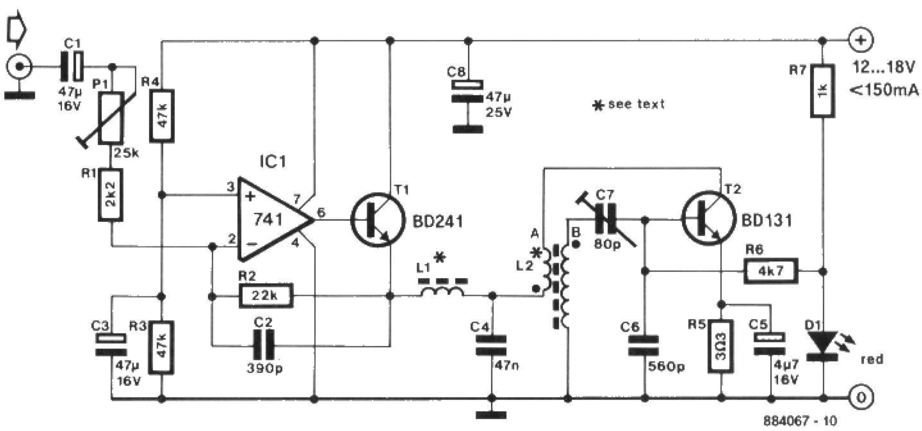


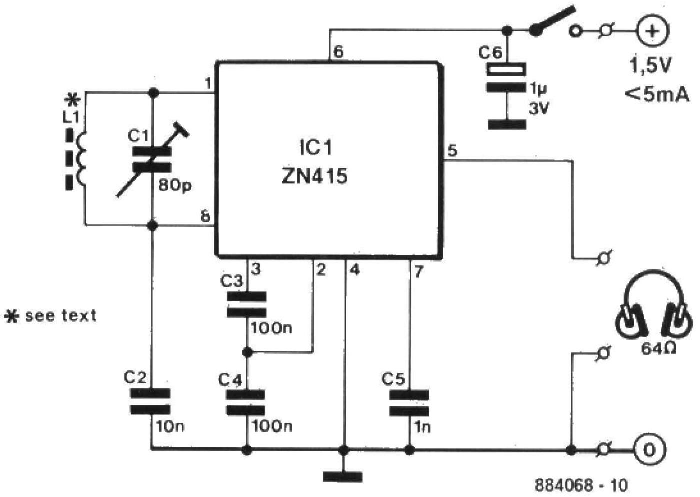
042 WIRELESS HEADPHONES (TRANSMITTER)

A circuit for the transmission, with good quality, of the sound output of a TV receiver over a couple of metres.
The input signal for the circuit is taken from the headphone or video recorder output of the TV receiver. If these are not available, NEVER ATTEMPT TO FIT ONE YOURSELF: THE CHASSIS OF THE TV SET MAY BE AT A LETHAL HIGH VOLTAGE.
The audio signal is amplified by IC_1, which has been given some extra 'body' by the addition of an output buffer, T_1. Capacitor C_4 has a potential that is equal to half the supply voltage (via R_3-R_4), on to which the amplified audio signal is superimposed. The resulting varying direct voltage is used as the supply voltage for emit transistor T_2 via the primary of L_2. The carrier oscillator, also formed by T2, can oscillate between 1,750 kHz and 3,500 kHz. The consequent amplitude-modulated signal across the secondary of L_2 is strong enough to span a few metres. A ferrite rod is used as transmit antenna.
Diode D_1 serves two functions: it indicates that the transmitter is 'on' and it stabilizes the direct voltage (about 1.5 V) for the oscillator. The supply voltage for the oscillator is thus independent of the 12—18 V line.
The inductors are easily made. A T50-2 toroid with 80 turns of 0.2 mm dia. enamelled copper wire is used for L_1. A ferrite rod of 10 to 20 cm is needed for L_2: L_{2A} consists of three turns of 0.6 mm dia. enamelled copper wire and should be wound at the ground side of L_{2B}. This secondary winding consists of 30 turns 0.5 mm dia. enamelled copper wire.
It is recommended to power the circuit from a mains adaptor, because a current of up to 150 mA may be drawn.



043 WIRELESS HEADPHONES (RECEIVER)

To arrive at a suitable headphone receiver that meets the requirements of being light, battery-powered, and offering good-quality reproduction, a Ferranti ZN415 was chosen.
This IC contains a complete AM detector, an output amplifier, and operates from a single 1.5 V battery.
The circuit shows the ZN415 in its standard application as a medium wave receiver. Circuit C_1—L_1 is, however, tuned to a frequency above the medium wave band. The output stage drives a high-impedance headphone without any problems. The circuit draws a current not greater than 5 mA, which ensures a good battery life.
The tuned circuit, C_1—L_1, receives the signal from the transmitter described in the preceding article. The inductor consists of 40 turns 0.2 mm dia. enamelled copper wire close-wound on a 20 mm dia. ferrite rod. For optimum reception, C_1 must be adjusted with a non-metal screwdriver. Note that the transmit frequency lies somewhere between 1,700 and 3,400 kHz.



044 LEAD-ACID-BATTERY CHARGER

Modern sealed lead-acid batteries are simplicity itself in use. In contrast to NiCd batteries, they may be charged by connecting them to a constant voltage (at the correct level). The charging current then gives a pretty good indication of the state of charge.
These batteries may also be charged at a rapid rate, as long as the charging current is limited at the onset of the charging process. Dependent on the make, a charging current of several times one tenth of the capacity in Ah is permissible. For instance, a 5 Ah battery may be charged with an onset charging current of 1 A. The charging voltage may then be 2.45 V per cell. At such a (relatively) high voltage, the current has to be limited, otherwise the onset charging current through a flat battery may be as high as 10 A.
The proposed charger, whose circuit is shown in Fig. 1, incorporates a 'standard' voltage regulator, IC_1, and a variable current limiter consisting of T_1, R_1, and R_4. As soon as the current through R_1 becomes too large, T_1 switches on and the output voltage drops. The output voltage is given by: $1.2(P_1+R_2+R_3)/R_3$ |volts|.
The current limiter becomes operative at a current of $0.6/R_1$ |amperes|.
The charging voltage for a 6-V battery that is required to be charged rapidly is $3\times2.45=7.35$ V. The total effective value of R_2+P_1 should then be 585 ohms. In practice, this value may be slightly different.
For charging 12-V batteries, the value of R_2+P_1 needs to be about 1290 ohms.
The input voltage should be not less than 3 V higher than the output voltage. The LM317 needs a heat sink, not because it is easily damaged, but because it cannot deliver its full output current at high temperatures.
It is, of course, possible to use the proposed circuit as a common supply unit.

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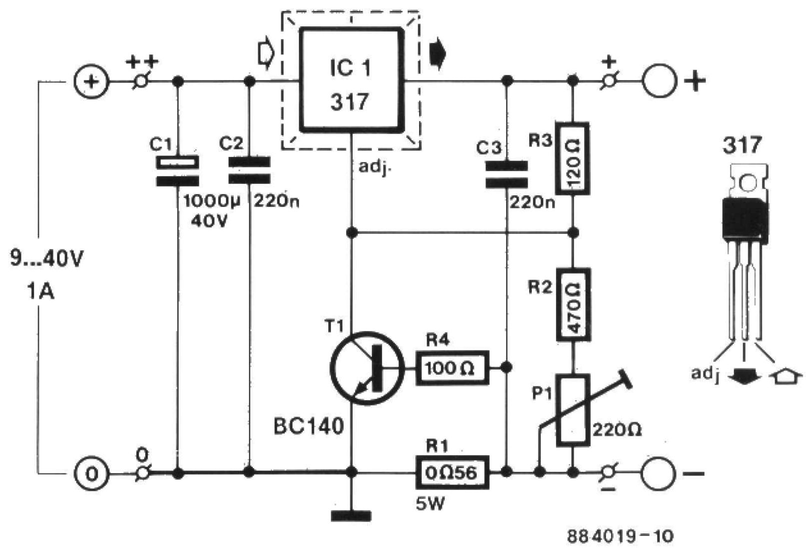

Parts list

Resistors (±5%):
R_1 = 0R56; 5 W
R_2 = 470R
R_3 = 120R
R_4 = 100R
P_1 = 220R preset

Capacitors:
C_1 = 1000μ; 40 V
C_2;C_3 = 220n; MKT

Semiconductors:
T_1 = BC141
IC_1 = LM317

Miscellaneous:
Heat-sink for IC_1.
PCB Type 884019 (not available through the Readers Services).

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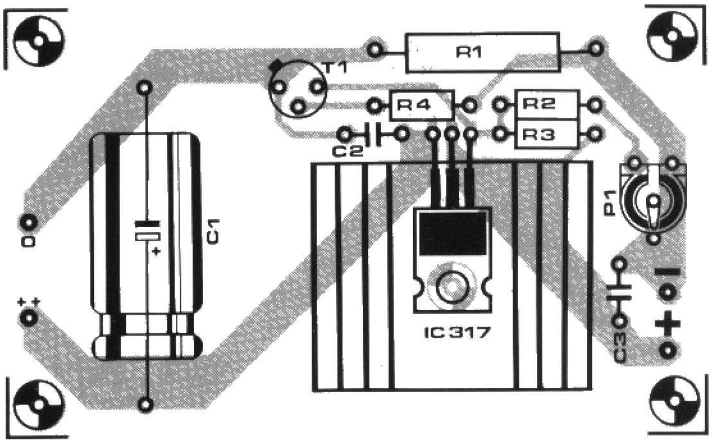


045 STEP-UP SWITCHING REGULATOR

Maxim Integrated Products have recently introduced a series of integrated step-up switching regulators designed for simple, minimum component count DC-DC converters. All control and stabilization functions are contained in an 8-pin DIP package: a bandgap voltage reference, oscillator, voltage comparator, catch diode, and an N-channel medium power MOSFET. In addition, the ICs have a built-in low-battery (LB) detection circuit.

One of these new chips is the Type MAX641, which is of particular interest for no-break 5 V supplies in computers. In the application shown here, the output current of the step-up regulator is boosted by an external bipolar power transistor, T_1. The low-battery detector compares the voltage at input LBI with the internal +1.31 V bandgap reference. Output LBO goes low when the voltage at pin 1 drops below 1.31 V. The low-battery threshold voltage, U_{LB}, is determined by potential divider R_1-R_2 as

$$U_{LB} = 1.31(R_1/R_2 + 1) \qquad [V]$$

R_2 is typically 100 kΩ. In the application circuit shown here, LED D_1 at the LBO output lights when the input voltage drops below 2.62 V.

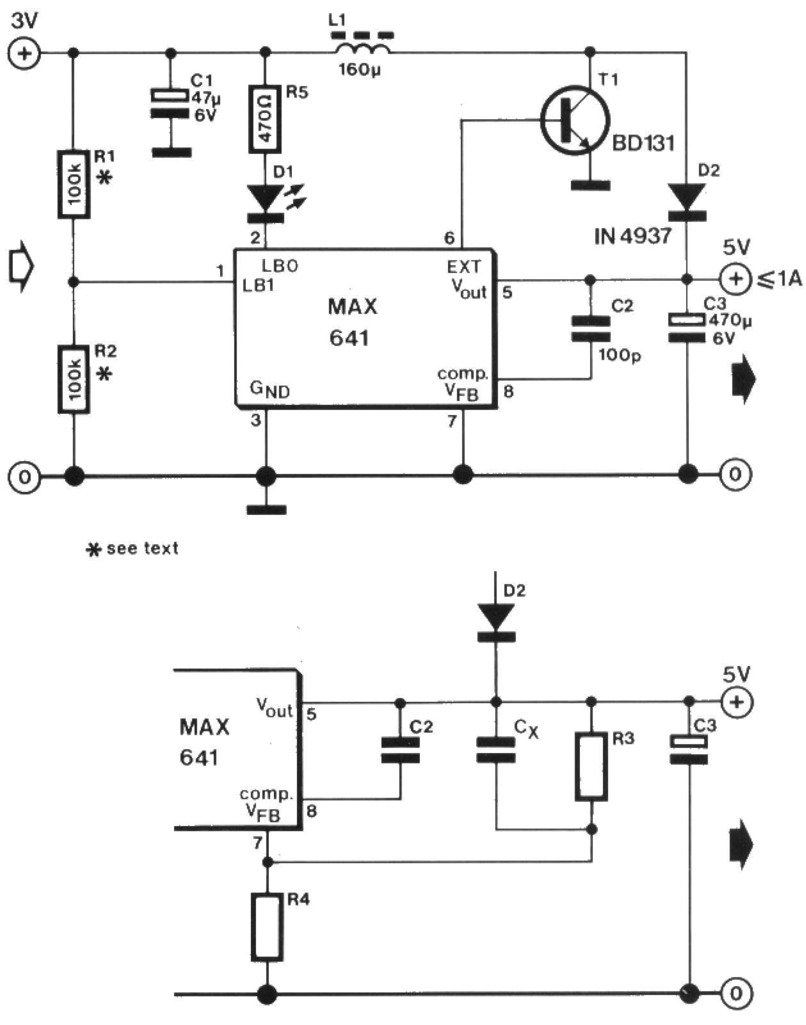
It is possible to make the output voltage adjustable by connecting input V_{FB} to a potential divider R_3-R_4 instead of ground. This option is shown inset in the circuit diagram. The output voltage, U_o, then becomes

$$U_o = 1.31(R_3/R_4 + 1) \qquad [V]$$

R_4 is, again, typically 100 kΩ. C_x is 100 pF. Remember to observe the voltage rating of C_3.

Maximum output current of the circuit is 1 A. The input voltage should remain below 5 V. Maximum conversion efficiency is about 80%.

As to components: the minimum value for L_1, L_{min}, is expressed by



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$$L_{min} = U_{in}/(2 f_0 I_{max})$$

I_{max} depends on the current rating of the inductor and external power transistor. Factor f_0 is the converter oscillation frequency, 45 kHz. The available output power can be increased by either raising the input voltage or lowering the inductance. This causes the current to rise at a faster rate, and results in a higher peak current at the end of each cycle. The available output power increases since it is proportional to the square of the inductor current. The calculation of

the maximum inductance of L_1 is, unfortunately, relatively complex, and falls outside the scope of this introduction to the MAX641. The inductor should be able to handle the required peak currents whilst having acceptable series resistance and core losses. The inductor in this application circuit should be rated at 2.5 A minimum.
Due account should be taken of the relatively high ripple amplitude at the output of the converter. The ripple voltage is composed of high (45 kHz) and low-frequency components, and is practically impossible to suppress further.
Finally, D_2 should be a fast Schottky diode. Alternatives to the type shown in the circuit diagram are the Types 1N5817 (1 A), 1N5821 (3 A), or the BYV27 (2 A). General purpose rectifiers from the 1N400x series are not recommended because their slow turn-on time results in excessive losses and poor efficiency.

Source: *Fixed Output 10 Watt CMOS Step-Up Switching Regulators.* Maxim Integrated Products.

046

FISHING AID

This circuit provides audible and visible warning when a fish is nibbling the bait. Although this event is fairly easy to signal with electronic means, the circuit is relatively extensive to ensure that it can be powered from a 9 V battery.
The circuit is based on a slotted opto-coupler Type CNY37, and a home made notched wheel. Unfortunately, the current amplification of slotted opto-couplers is very low (0.02 min.), requiring considerable current to be fed through the LED before a usable collector current flows in the phototransistor. To avoid rapidly exhausting the battery, MMV_1 pulses the LED at about 250 Hz and a duty factor of 0.05. MMV_2 detects the presence of these pulses. When a fish pulls at the bait, the notched wheel revolves in the slot, and intermittent pulse bursts are received at the trigger input of MMV_2. Green LED D_1 lights, buzzer Bz sounds, and bistable N_3-N_4 is set, so that red LED D_2 flashes at a 1.5 Hz rate. D_1 and the buzzer are turned off when the fish gets off after nibbling the bait, but D_2 continues to flash. The circuit around N_1, T_2 and C_3 then serves to keep the current consumption as low as possible. The circuit can be reset by pressing S_1.
Preset P_1 enables adjusting the frequency of the buzzer oscillator between 600 and 2500 Hz. When several fishing-rods are being used, each can be assigned a particular signal tone. The buzzer can be switched off by means of S_2.
A suggested construction of the light barrier and the notched wheel is shown in Fig. 2. A small shaft is used in combination with a reel around which the fishing line revolves. The slots cut into the detection wheel should not be too wide: 1 mm is a good starting value. The detection sensitivity is determined by the number of slots in combination with the reel diameter. The light barrier should be screened from daylight.
In the stand-by condition, the circuit consumes no more than 4 mA, which goes mainly on account of the LED in the opto-coupler. In the actuated state, the current consumption rises to about 12 mA.

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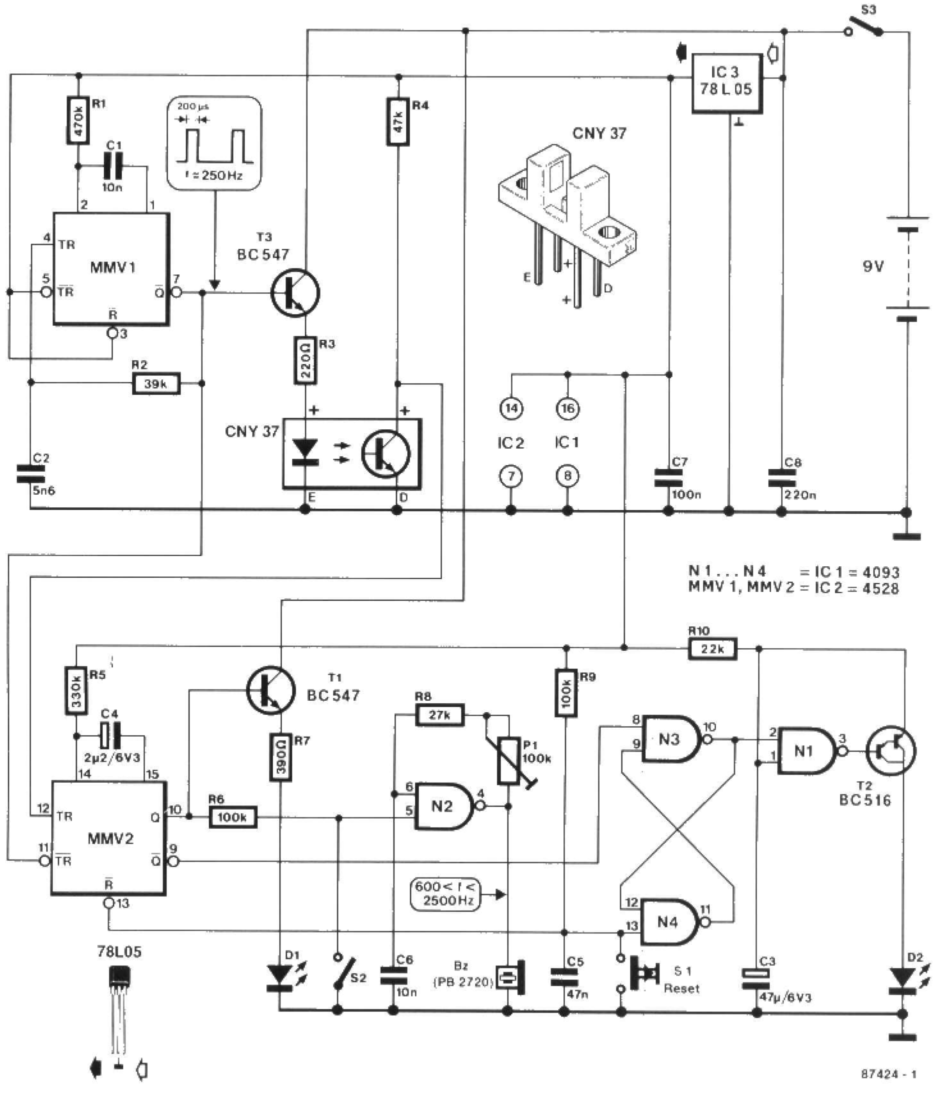

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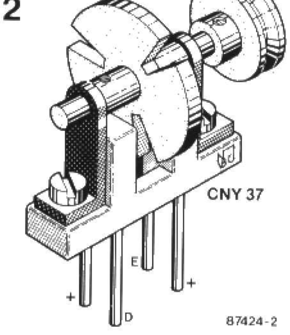
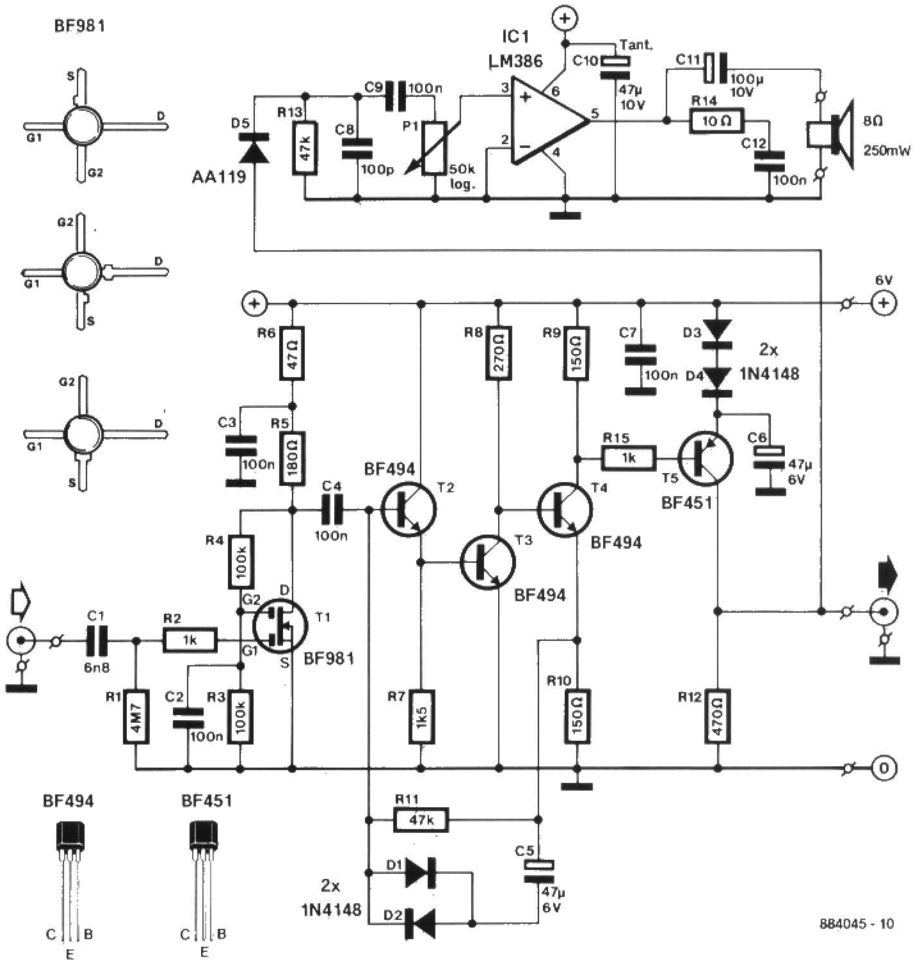


047

Wideband RF signal tracer

This simple and versatile circuit can aid in troubleshooting defective RF amplifier circuits. The usable frequency range of the tracer is about 100 kHz to 30 MHz. Measured signals (0.5 mV to 500 mV) are amplified, detected and made audible with the aid of a small loudspeaker.
MOSFET T_1 functions as an amplifier with a high input impedance to avoid loading the signal source. Transistors T_2, T_3 and T_4 form a high-gain logarithmic amplifier that drives AM demodulator T_5-D_5. A single chip AF power amplifier, IC_1, is included to make detected signals audible. Testing of RF equipment is carried out simply by "probing around" at suitable locations in the circuit and listening to the detected signal, whose relative amplitude can provide an indication of possible sources of malfunction. The tracer's logarithmic amplifier obviates the need for frequent re-adjustment of the volume control, P_1. The unit is so sensitive that it produces audible output when the input is only held near the circuit section under test.
As to construction of the tracer, this is best fitted in a short length of ABS tubing to make a probe with three connecting wires for the supply voltage and the loudspeaker. Constructors are advised to strive for ample RF decoupling and short connections in view of the relatively large bandwidth. Current consumption of the tracer is about 100 mA from a regulated 6 V supply.



048

DRIVER FOR BIPOLAR STEPPER MOTORS

For some applications, the *Universal control for stepper motors* (see [1]) may be considered too extensive a circuit. Many small motors with limited speed range can be equally well controlled by a relatively simple circuit, based on, for instance, the Type SAA1027 or TEA1012 [2]. Most commercially available controllers are, however, intended for driving unipolar stepper motors, which are now gradually superseded by bipolar types of similar size. In practice, the latter can provide a larger torque, but require a different type of controller.
The recently introduced Type MC3479P from Motorola requires a minimum of external components for controlling a bipolar stepper motor. The maximum quiescent stator current, I_s, depends on the value of resistor R between pin 6 and ground:

$$I_s = (U_b - 0.7)/0.86R \text{ [mA]}$$

where R is given in kΩ. The above relation between I_s and R is valid as long as the output transistors are not operated in the saturated area. The saturation point is reached sooner at low levels of the supply voltage, or when the ohmic resistance of the stator winding is fairly high. The manufacturers state a maximum current of 350 mA per stator.
The supply voltage for the motor (pin 16) depends on the ohmic resistance of the stator windings, and is allowed to vary between 7.2 and 16.5 V. When a high supply voltage is used, it must be remembered that the output transistors will not operate in the saturated area to prevent exceeding the set stator current, I_s. The current control used here allows a fairly high step rate at the cost of an increase in the dissipation of the driver IC, particularly when the motor is held stationary. If necessary, the MC3479P can be cooled by connecting the 4 central ground terminals to a relatively large copper surface on the PCB.
The integrated controller has 4 TTL and CMOS compatible inputs (see Fig. 1):

CLK (pin 7): every rising edge of the clock signal causes the motor to revolve one full or one half step, depending on the level at pin 9. The maximum step rate and the minimum pulse width are 50 kHz and 10 µs respectively.
CW/CCW (pin 10): the logic level applied here determines the motor's direction of travel.
$\overline{\text{F}}$/H step (pin 9): this input allows selection between full (0) or half step (1) operation—see Fig. 3.

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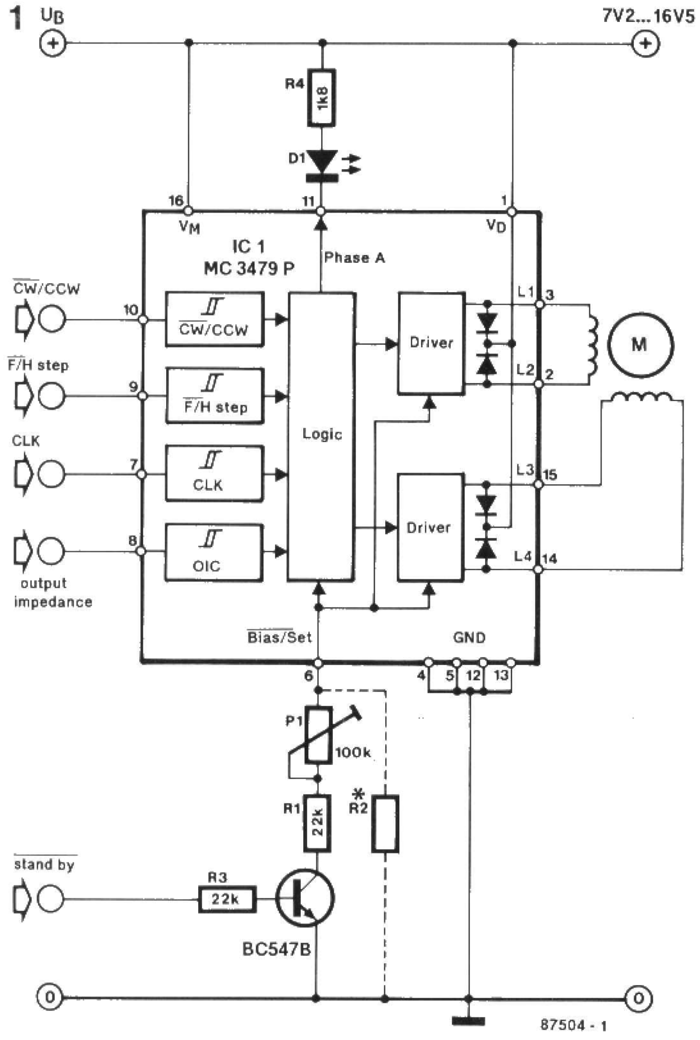

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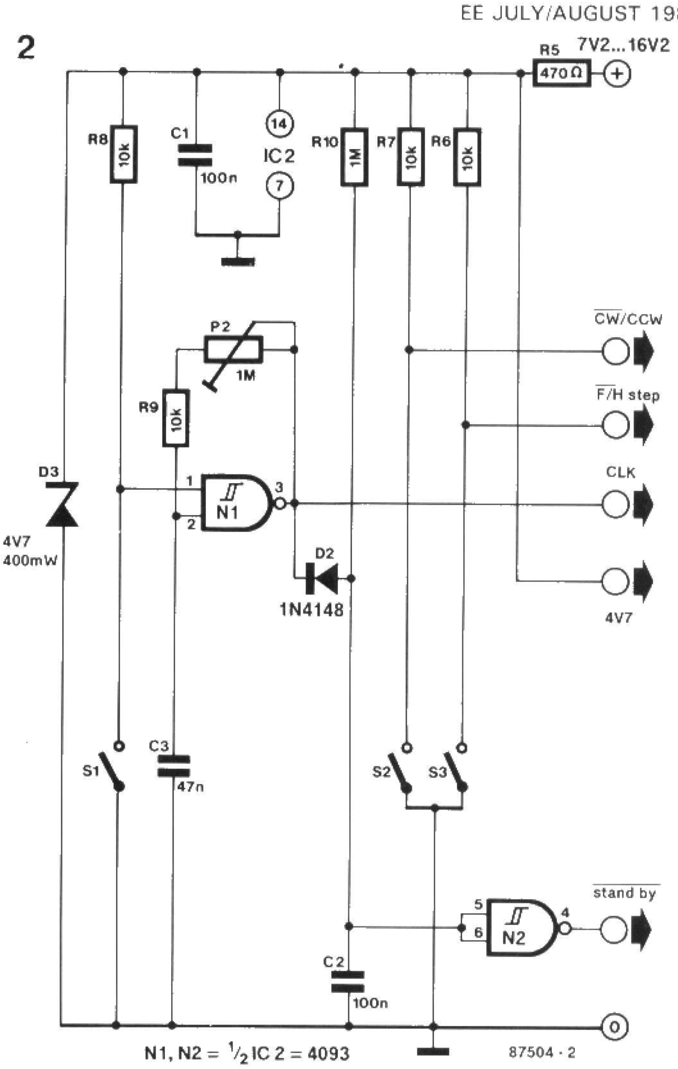

OI (pin 8): this output impedance selection input is only effective in the half step mode. It determines whether the stator winding is effectively disconnected from the driver (0), or connected to the positive supply at both ends (1). The latter option improves the damping of the motor in the half step mode, and will prove useful at relatively low step rates.

Pin 11 of the driver IC is an open-collector output with a current capacity of 8 mA, activated during period A in Fig. 3. A LED connected to this output will flash rhytmically when the motor is running.

Transistor T_1 was added to obtain a reset function. No stator current flows, and the logic circuitry in the driver is reset, when the stand by input is driven low. When a logic 1 is applied, the motor is energized starting from state A. The addition of R_2 makes it possible to switch the driver to the power-down state, rather than the reset state. The stator current is reduced to the value set with R_2, as shown in the above formula.

The motor driver is probably best controlled by a computer output port. The circuit in Fig. 2 is intended for stand-alone applications. It is composed of a supply, R_5-D_3, an oscillator, N_1-C_3-R_9-P_2, and a re-triggerable monostable multivibrator, N_2-C_2-R_{10}-D_2. When S_1 is opened, the oscillator is enabled, and the motor

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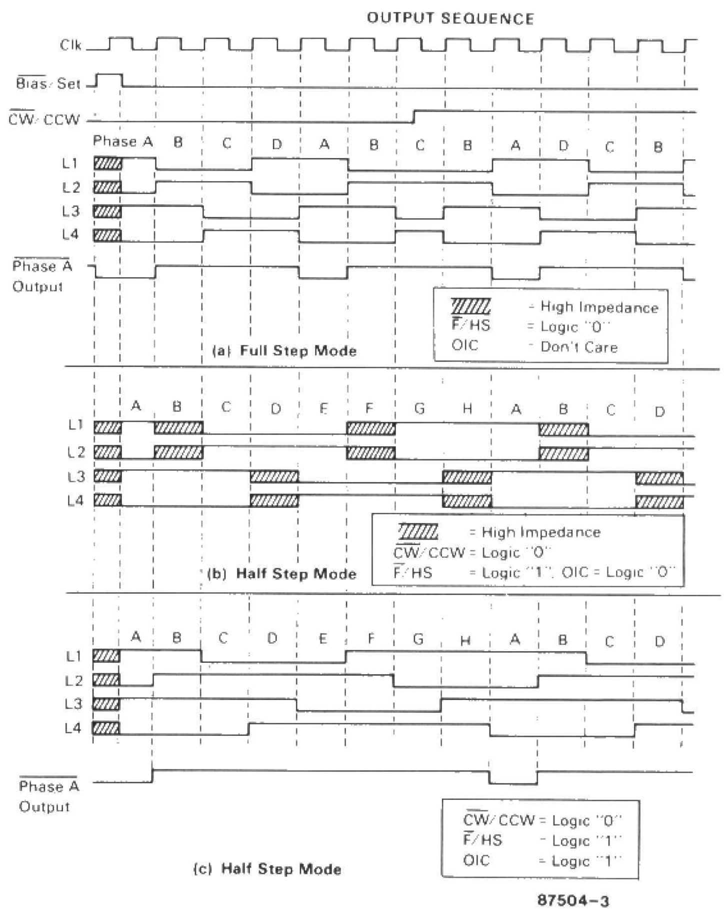

will start running. The clock frequency, i.e., the step rate, is adjustable with P_2. The monostable will remain set via D_2, and T_1 will conduct, as long as clock pulses are applied to the motor driver. The amount of ever reversing stator current is limited by the stator inductance, but can still be increased with the aid of P_1. When the motor stops, T_1 is turned off, and the stationary stator current is reduced to the value set with R_2. The above arangement keeps the dissipation of the motor and the driver within reasonable limits.
The current consumption of the complete circuit is practically that of the motor alone (700 mA max.). The motor driver IC consumes about 70 mA.

References:

[1] *Universal control for stepper motors.* Elektor Eelectronics, January 1987.
[2] *Stepper motor control.* Elektor Electronics, July/August 1986.

049 NON-INTERLACED PICTURE FOR ELECTRON

Owners of the Acorn Electron home computer may well object to its interlaced, and therefore slightly instable, picture. There is a trace of display flicker in non-moving areas on the screen, and this is mainly due to the internal video procesing circuitry operating on the basis of interlacing, a technique used in conventional TV transmission for smoothing the appearance of moving picture areas. Arguably, interlacing is not very useful in computers, since these work with text in most applications. Special displays with a relatively long afterglow time are no remedy for this awkward problem, and that is why the present circuit was designed. It effectively switches off the interlace function, and so ensures a restful display, albeit that the individual lines that make up the characters become slightly more prominent.
Figure 1 shows that a TV picture is composed of 625 lines divided between 2 rasters of 312.5 lines each. In an interlaced picture, these rasters are vertically shifted by one line. This is done by starting the second raster x and a half time later than the first raster. Interlacing can thus be rendered ineffective by starting the second raster half a line period earlier (i.e., after 312 lines rather than 312.5). To retain the normal number of lines (625), the second raster is arranged to comprise 313 lines.
The ULA chip (Uncommitted Logic Array) in the Electron computer provides a horizontal and a composite synchronization signal, which are shown in Figs. 3a (HS) and 3b (CSYNC) respectively. With reference to Fig. 3c, and the circuit diagram in Fig. 2, MMV_1 forms a new vertical synchronization pulse, VS, with the aid of the CSYNC signal. The period of pulse VS is different for the first and second raster, so that MMV_2 is needed to make VSYNC equally long in both. MMV_2 is triggered on the first line pulse (HS) that occurs when VS is active, and is retriggered when VS goes low—see Fig. 3d. The length of the VSYNC pulse so made is about 160 μs, or about 2.5 times the line time (64 μs). The HS and the new VS signal are combined in XOR gate N_2 for driving the video modulator. Gate N_1 serves to buffer the HS output of the ULA.
The final results obtained with the circuit depend mainly on the type of TV

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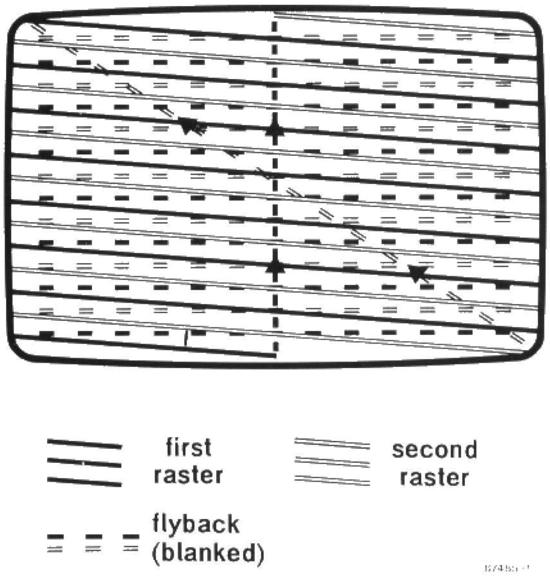

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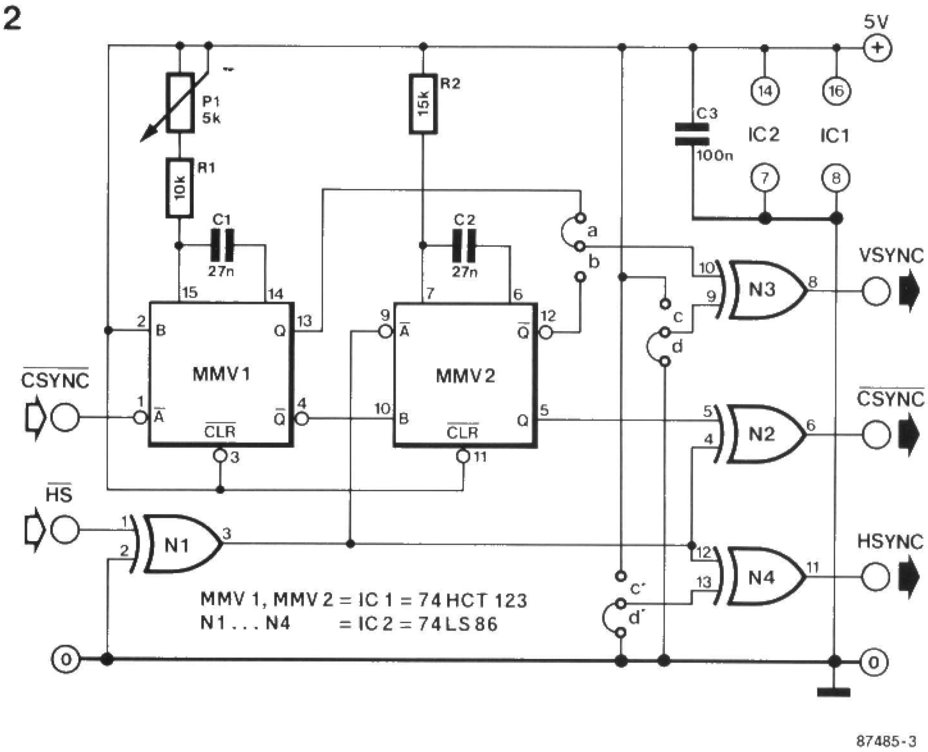

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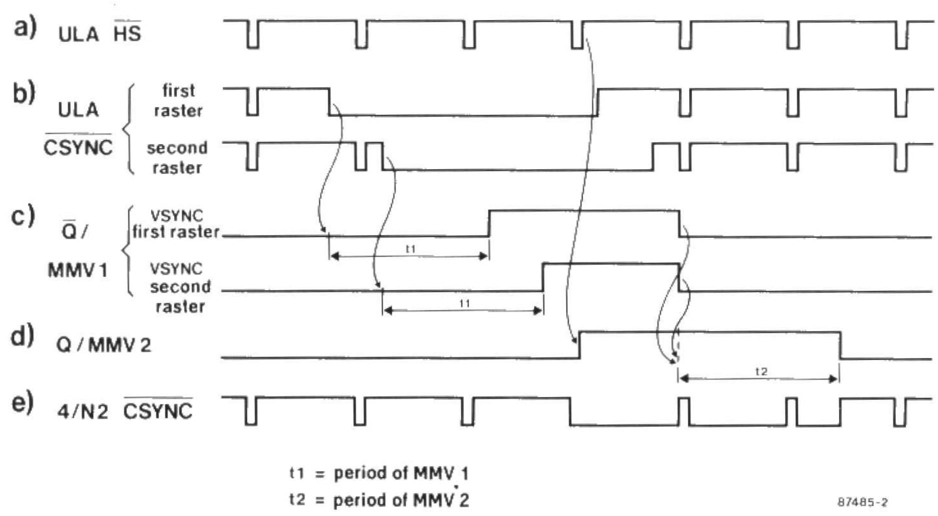

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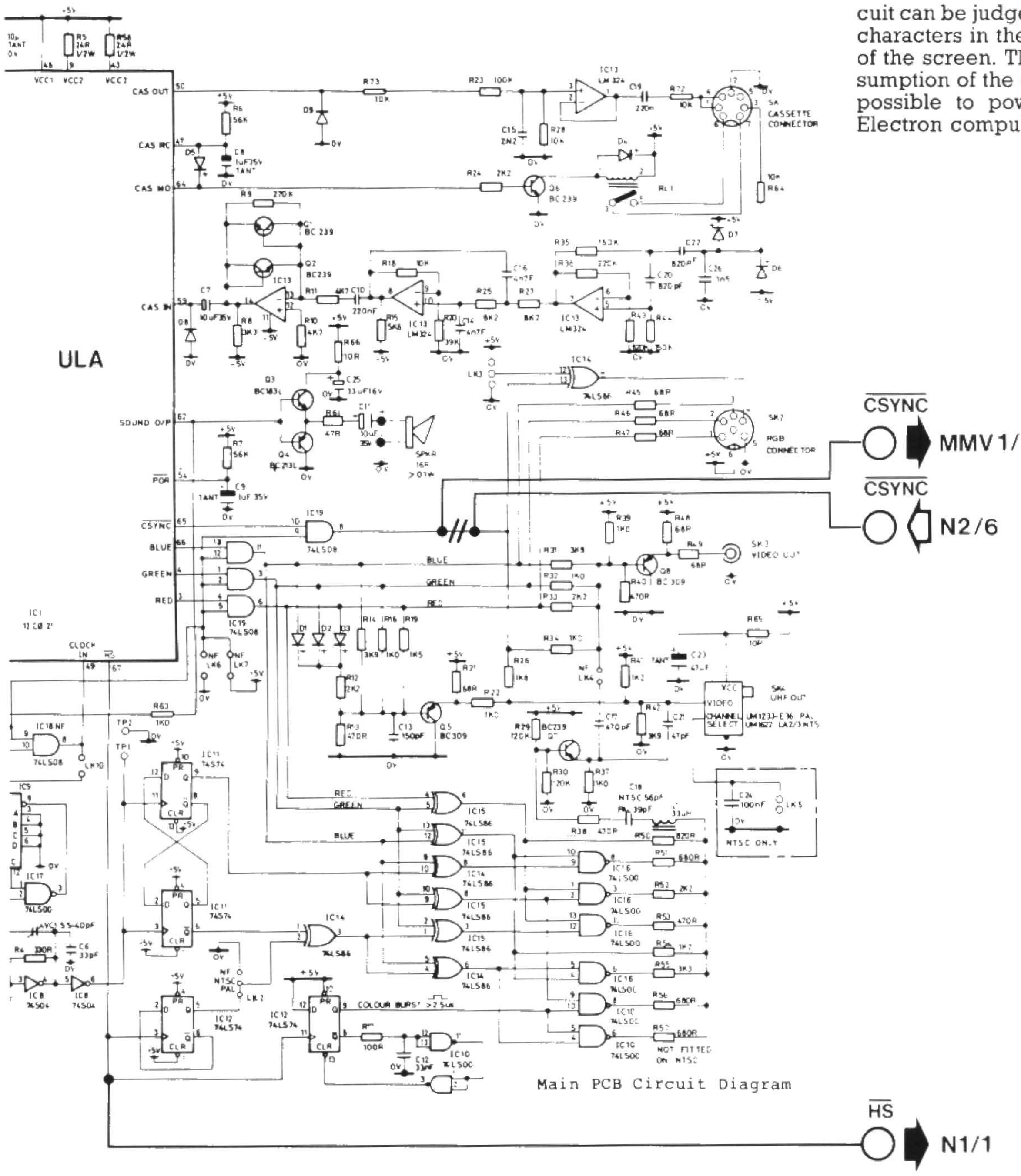

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set or display used, and may not be optimum when the TV is driven via its RF input. On an older type monochrome set, the central area of the picture was stable, but the upper and lower areas gave a less favourable look. Good results were obtained, however, from the use of Type TX chassis, which are currently the basis of TV sets sold under many different names and licenses. Even better performance can be expected from a video monitor, whose (TTL compatible) H and V synchronization inputs can be driven by N_4 and N_3 respectively. The polarity of the sync signals can be selected with the aid of wire jumpers. Connections c and c' result in $\overline{VSYNC}$ and $\overline{HSYNC}$. The choice between jumper a or b depends on the type of display used. Preset P_1 is adjusted until the picture appears vertically synchronized: the adjustment is fairly critical when jumper a is used. The final results obtained with the circuit can be judged from looking at a few characters in the upper and lower area of the screen. The modest current consumption of the circuit, 10 mA, makes it possible to power it direct from the Electron computer.

050 FRUIT MACHINE

This is one of the very few 'one-armed bandits' to which the maxim *the sole way to win is not to gamble* is not applicable. In other words, this circuit does not have a slot for inserting coins: every play is free.
Actuation and release of the 'PLAY' button, S_1, causes the circuit to become operative. Series regulator T_1 is driven into saturation by T_2, which is controlled by N_2-N_7. The outputs of N_2, N_{11}, and N_{10} go high in succession, and disable counters IC_5, IC_4 and IC_3, which are all clocked by oscillator N_{12}-N_9, and reset by the pulse at their Q_3 output. The 3 LEDs driven by each of the counters, therefore, lights cyclically. When a counter is disabled by the high level at its $\overline{CE}$ input, one of the LEDs in the 3 groups remains illuminated. The output state of the counters is not predictable because of the inconstant delay between the disable instants. NAND gates N_{13}-N_{15} detect the winning combinations, i.e., LED D_2 lights, and Bz_1 is sounded, when 3 identical counter outputs are activated. Note that diodes D_3-D_5 form a 3-input OR gate, and that the buzzer also produces sound when the LEDs are flashing, since the pulses at output Q_2 of IC_3 enable the oscillator intermittently.
The play is ended when the voltage across C_3 is high enough for gate N_7 to change state. T_2 is turned off, and T_1 no longer powers the circuit. An on/off switch is not required for the fruit machine, thanks to its very low current consumption in the de-activated state.

Parts list

Resistors (±5%):
R1;R3;R4;R5;R8;R10 = 10K
R2 = 4K7
R6 = 220K
R7 = 1M5
R9 = 470K
R11 = 3M3
R12;R14 = 33K
R13 = 100K
R15 = 1K0
R16 = 330R

Capacitors:
C1;C3 = 4μ7; 16 V; axial
C2;C5;C6 = 10n
C4 = 1μ; 16 V; axial
C7;C8;C9 = 100n

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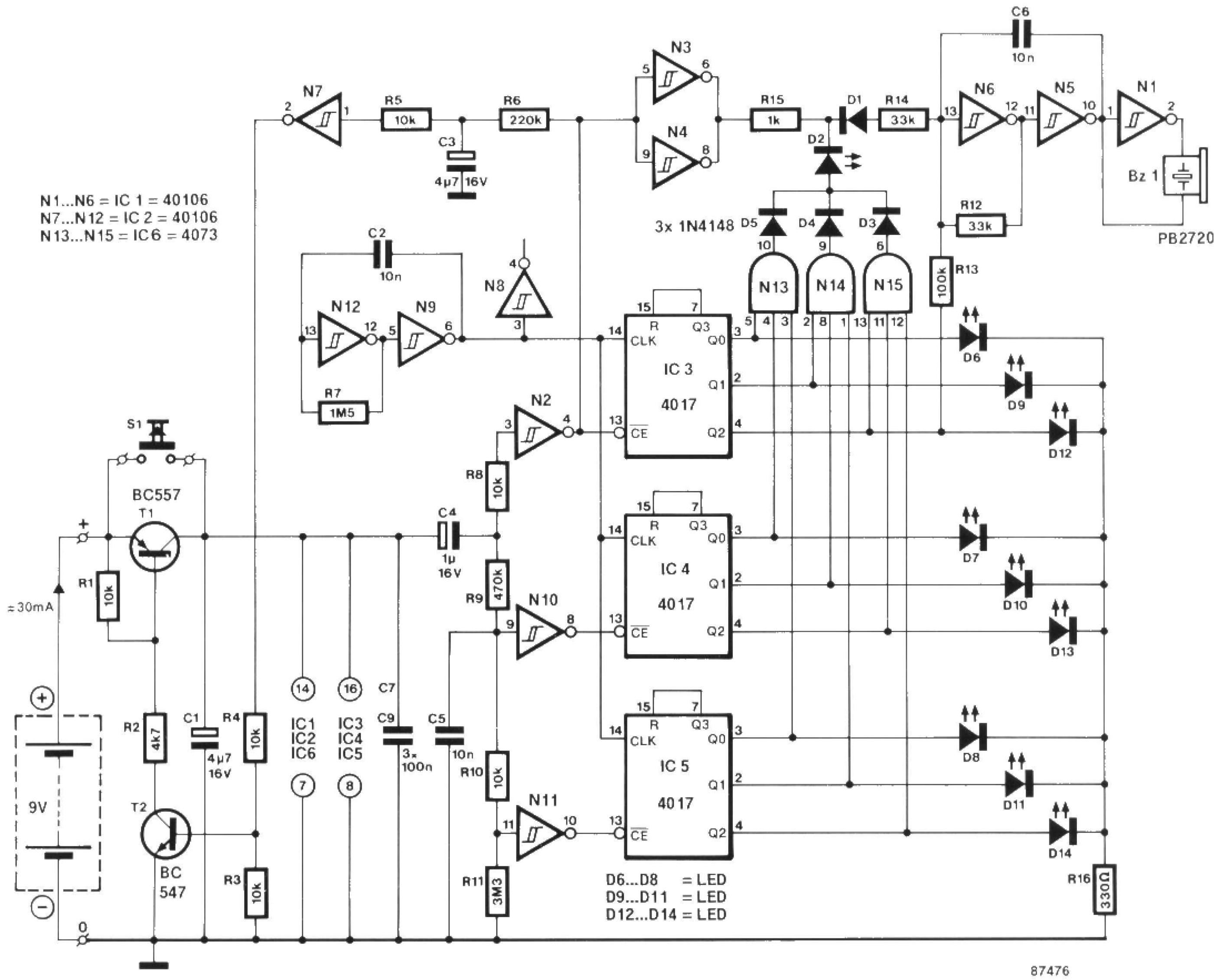

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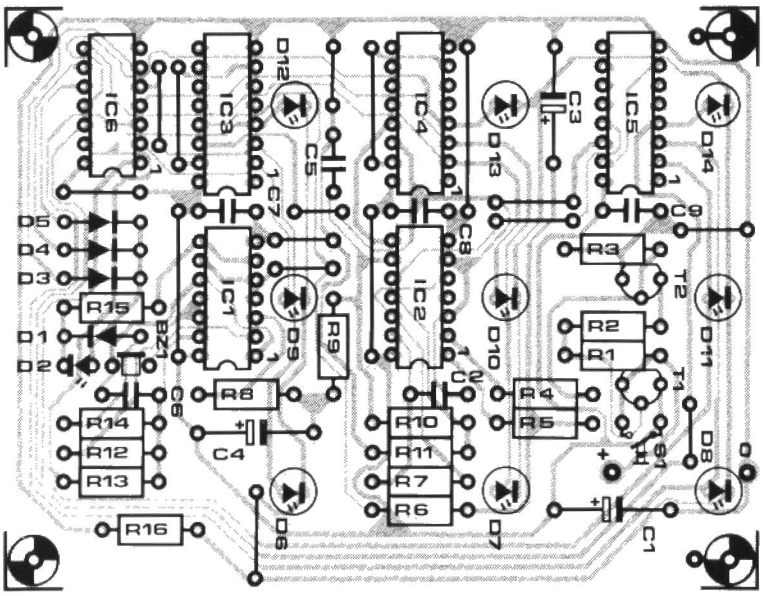


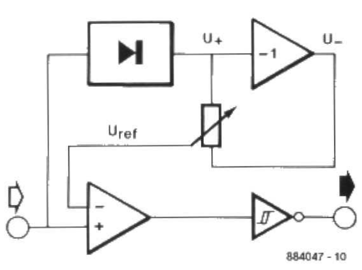
Semiconductors:
D1;D3;D4;D5 = 1N4148
D2;D6;D7;D8 = LED (red)
D9;D10;D11 = LED (yellow)
D12;D13;D14 = LED (green)
IC1;IC2 = 40106
IC3;IC4;IC5 = 4017
IC6 = 4073
T1 = BC557
T2 = BC547

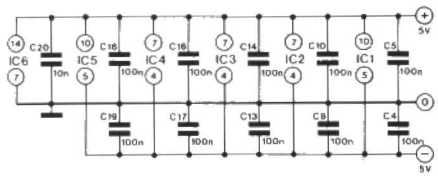
Miscellaneous:
S1 = momentary action push button.
Bz = PB2720 buzzer (Cirkit stock no. 43-27201).
PCB Type 87476 (available through the Readers services).

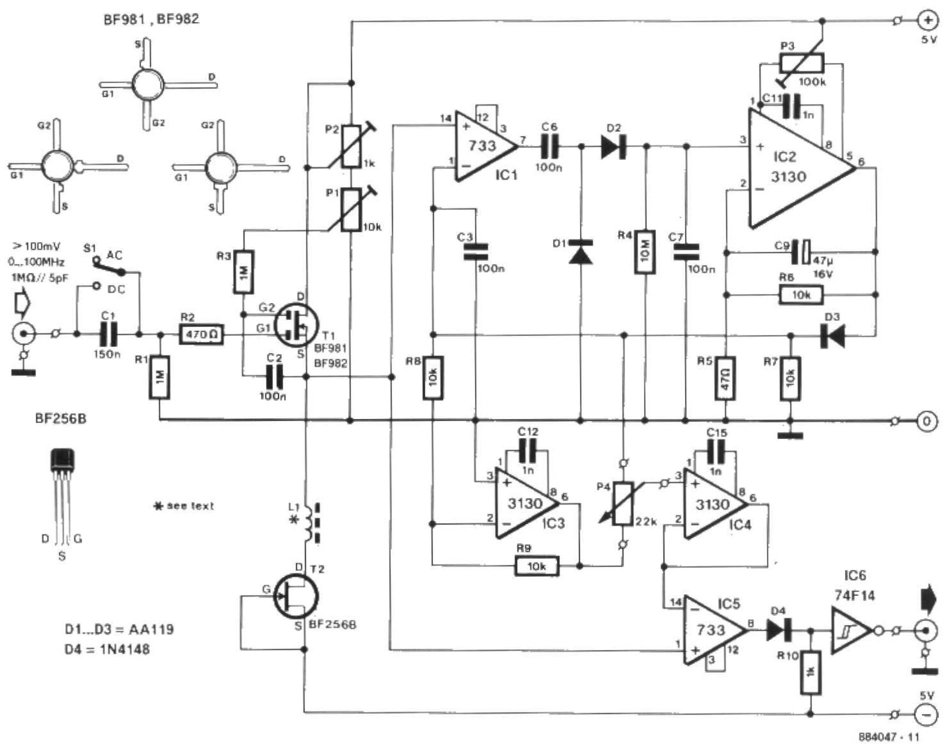
051 WIDEBAND LEVEL-INDEPENDENT TRIGGER PREAMPLIFIER

This circuit eliminates the difficulty in re-adjusting the trigger level of an oscilloscope or frequency meter any time the amplitude of the input signal changes. The block diagram shows that the trigger pulses are supplied by a fast comparator that compares the instantaneous input signal amplitude with a reference voltage deduced from the difference between the peak amplitude of the positive and negative half cycles of the rectified input signal. The circuit is fast enough to handle input signals with a frequency of up to 100 MHz, and has a sensitivity of 100 mV_{pp}.

With reference to the circuit diagram, the input signal is raised in a wideband preamplifier based around a UHF dual-gate MOSFET, T_1, fed by constant current source T_2. Presets P_1 and P_2 define the potential at the source of T_1, and hence form the fine and coarse offset compensation adjustments for the direct-coupled chain of opamps IC_1-IC_2-IC_3. The signal rectifier and direct voltage amplifier are formed by D_1-D_2-R_4-C_7 and IC_2. The relatively weak signal is raised further in direct-coupled opamps IC_3 and IC_4 for comparison with the amplified measuring signal in opamp IC_5. Schmitt-trigger/inverter IC_6 cleans the trigger signal before it is applied to the test instrument. The trigger sensitivity is set by potentiometer P_4. Choke L_1 is wound as 4 turns of 0.2 mm dia. enamelled copper wire through a small ferrite bead. MOSFET T_1 may be replaced by a Type BF991 or BF966 if either of these is easier to obtain locally. The circuit should be constructed with due attention paid to the relatively high frequencies it can handle. In this context, it is recommended to use a large copper area as an effective ground plane onto which the parts are fitted. The shortest possible connections, ample screening, and effective decoupling







of the supply voltage at various point in the circuit are also a must to ensure correct operation.
Optimum sensitivity is achieved by adjusting P_1, P_2 and P_3 for lowest offset measured at the output of IC_3. These adjustments are carried out after a warming-up period of a few minutes, and with the input of the preamplifier temporarily short-circuited.

052 FAST STARTING WIPER DELAY

A wiper delay is essentially a bistable multivibrator whose off-time is adjustable with a potentiometer. Many wiper delay circuits are based on the Type 555 timer in its standard application circuit, which has the disadvantage of introducing a delay of about 1.6 times the set interval before the first wiper action takes place. This is especially annoying when an interval of, say, ten or more seconds has been set. This circuit is also 555 based, but is unique in that it arranges for the wipers to be activated immediately at power-on.
The circuit diagram of Fig. 1 shows the internal organization of the 555 timer to aid in clarifying the operation of the present circuit. When S1 is closed, pin 6 is immediately pulled to +12 V because C_1 is discharged as yet (see also Fig. 2b). The bistable in the 555 is reset, the output goes low, and Re_1 is energized. This forms the basic difference with the standard application of the 555, where C_1, connected as shown in Fig. 2a, delays the relay action until charged to ⅔ of the supply voltage. Returning to Fig. 1, C_1 is charged via R_2 and the 555's internal transistor when the output is activated. The bistable is reset when the voltage at pin 2 drops below ⅓Vcc, causing the relay to be de-energized, and C_1 to be discharged via R_1-P_1. The discharge time, and hence the wipe interval, is defined by the setting of P_1. When this is set to the shortest delay, the wiper motor is constantly powered via Re_1, since C_1 is not charged via P_1-R_2 only, but effectively via voltage divider P_1-R_1-R_2 also. The wiper delay is fed from the 12 V car battery, and its current consumption is practically that of the relay alone. Note that the coil current may not exceed 200 mA.

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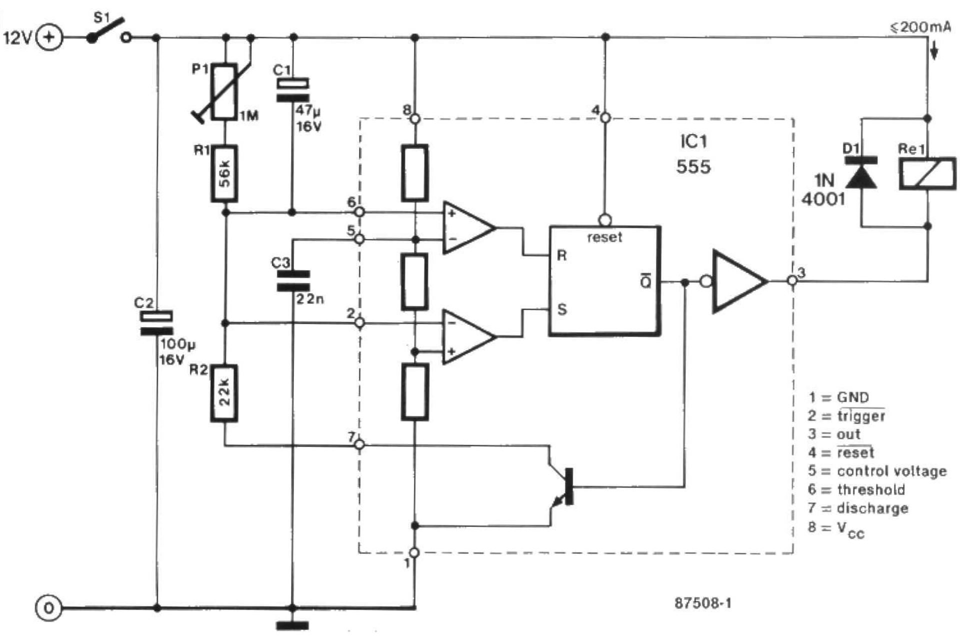

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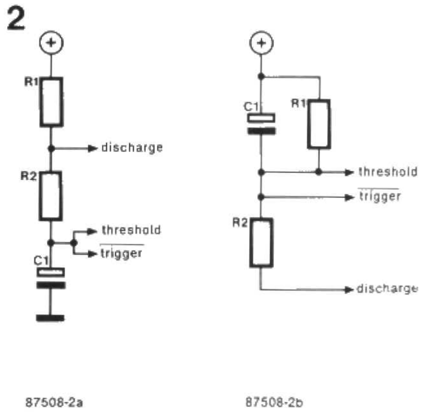
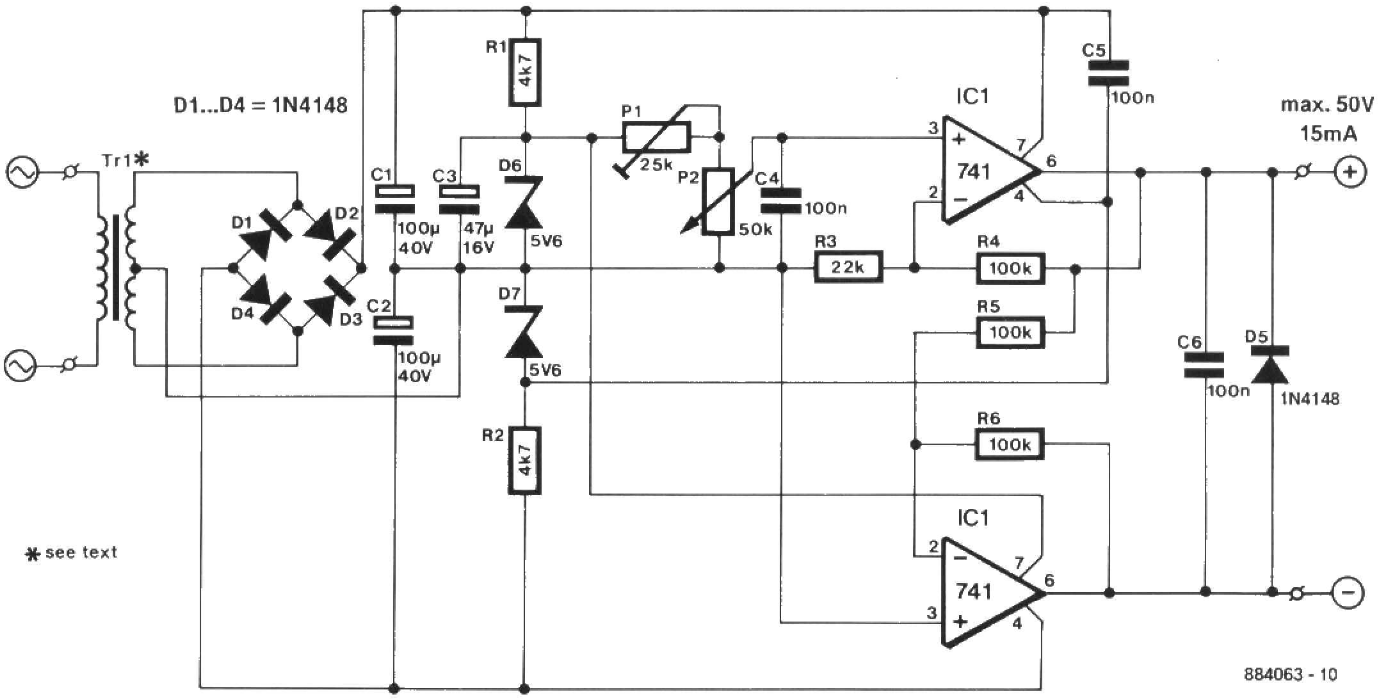


053 TEST-VOLTAGE SUPPLY

For testing zener diodes, base-emitter breakdown, diacs, and so on, a fairly high voltage is needed. The usual type of laboratory power supply is not suitable, because its output is normally of the order of only about 30 V. If the required current does not exceed 10 to 15 mA, it is possible to make a short-circuit-proof power supply with variable output voltage from 0 to 50 V from a handful of components as shown in the accompanying diagram.
Circuit IC_1 amplifies a direct voltage set by P_2 by a factor of about 6. Its output voltage should be about 25 V with respect to junction C_1-C_2. This voltage is inverted by IC_2, whose output is thus −25 V. There is then available either a symmetrical ±25 V potential with respect to junction C_1-C_2, or 50 V asymmetrical across the outputs of the ICs. The actual value of the voltage is set with P_1.
The maximum current is limited by the ICs to about 20 mA, so that the likelihood of damage to a component under test is very small. The output is

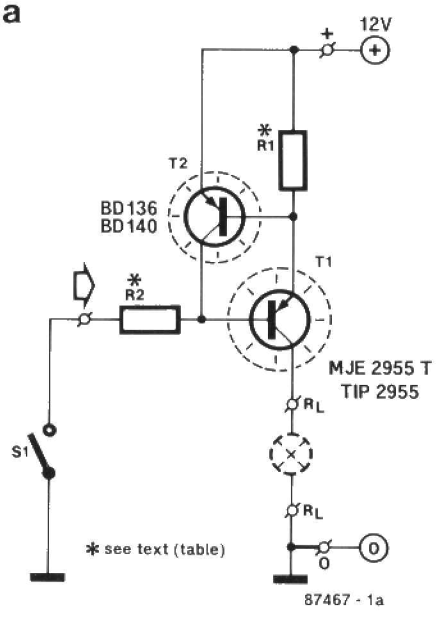


short-circuit-proof for an indefinite period.
To avoid common-mode problems, and also to make it possible to vary the output voltage to 0, the supply voltages to IC_1 and IC_2 overlap to some extent, which is arranged by D_6 and D_7. Zener D_6 also functions as the voltage reference. The supply to IC_1 must be decoupled separately by a 100 n capacitor; that to IC_2 is decoupled adequately by C_2 and C_3.
The mains transformer may conveniently (and inexpensively) consist of two 18 V types, otherwise a single 36 V unit is required. The secondary must be able to provide a current of 20 to 30 mA. If two transformers in series are used, make sure that they are in phase.
Before inserting the ICs into their sockets, check the voltage at pins 4 and 7: this should be not higher than 36 V if a 741C is used, or 44 V for other types (741A, 741E, and 741). If the voltage is too high, a transformer with a lower rated secondary (2 × 15 V or 30 V) should be used. If, however, the voltage at pins 4 and 7 becomes lower than 27 V, it may be impossible to obtain an output voltage of 50 V.

054 POWER SWITCH FOR CARS

Motorist are generally well aware that car fuses do not blow just like that. None the less, when something appears to be amiss in the electrical circuit, a new fuse is nearly always fitted prior to investigating the possible cause for the malfunction, which then, of course, costs two fuses. The circuits shown here are short circuit proof power switches, or electronic fuses with switch control dimensioned for relatively heavy (lamp) loads in a car. Both circuits are composed of a power switch, T_1, and a current limiter, T_2. The circuit is fully short-circuit and overload resistant, provided T_1 is adequately cooled, and the whole unit is constructed in a sturdy enclosure. The circuit in Fig. 1a has the lower voltage drop of the two, while that in Fig. 1b is used when a TO-218 style Type MJE2955T or TIP2955 is not obtainable. It is interesting to note that the plastic TO-218 package is mechanically interchangeable with the wellknown TO-3 outline, and enables ready mounting of

1a



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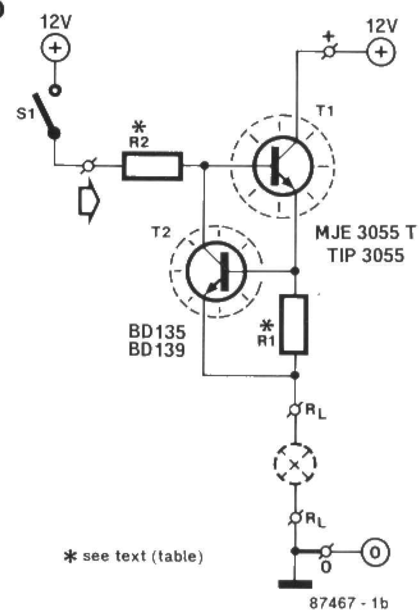

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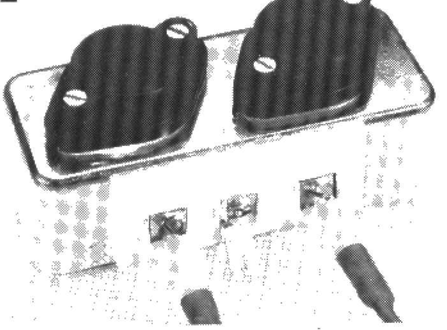

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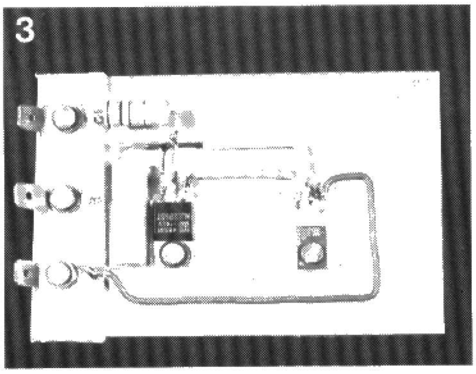

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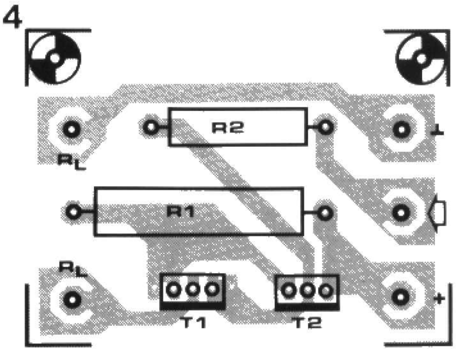


Table 1

Application	RL [W]	I [A]	R_1 [Ω]	R_2 [Ω]	Cooling T_1-T_2
Dashboard lighting	1	0.08	5.6	3300	not required
Courtesy light	2	0.17	2.7	1500	not required
Rear light or parking light	5	0.42	1.2	680	25 cm²
Brake light	18	1.5	0.33 (5 W)	180 (1 W)	225 cm²
Fog light or trafficator	21	1.75	0.27 (5 W)	150 (1 W)	225 cm²

Part list

Resistors (±5%):
R_1;R_2 = see text

Semiconductors:
T_1 = MJE2955T or TIP2955 (Fig. 1a)
T_1 = MJE3055T or TIP3055 (Fig. 1b)
T_2 = BD136 or BD140 (Fig. 1a).
T_2 = BD135 or BD139 (Fig. 1b).

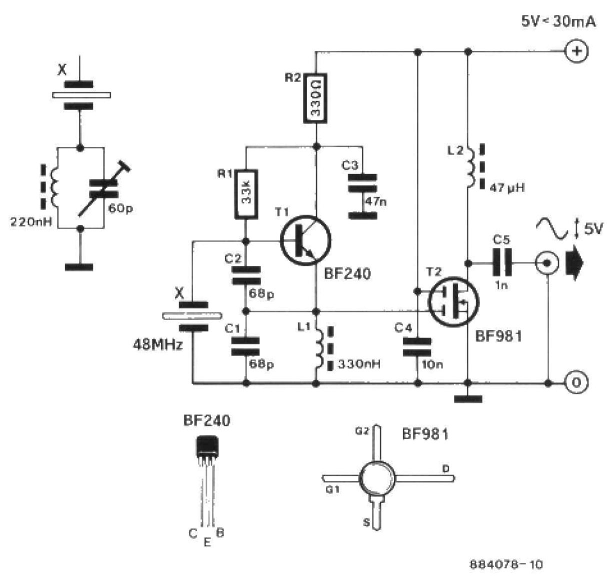
Miscellaneous:
S_1 = see text
PCB Type 87467 (not available through the Readers Services).

the transistor onto a flat surface using an insulating washer—see Fig. 2. The use of a die-cast enclosure and TO-3 style transistors is illustrated in Fig. 3. This unit houses two power switches, one of which has its contacts at the rear side. Pay great attention to the correct electrical insulation between the transistors and the enclosure, and, if required, that between the enclosure and the car body. Switch S_1 is the existing control for the relevant lamp in or on the vehicle. Note the difference in respect of the connection of S_1 in Fig. 1a and 1b. Table 1 shows how R_1 and R_2 are dimensioned in accordance with the current requirement of the load, and also gives a suggested area of the cooling surface. Finally, when the printed circuit board is used, T_1 should be a TIP2955 or a MJE2955T, not a MJE2955, since this has its outer terminals (B-E) reversed.

0 5 5 48 MHz CLOCK GENERATOR

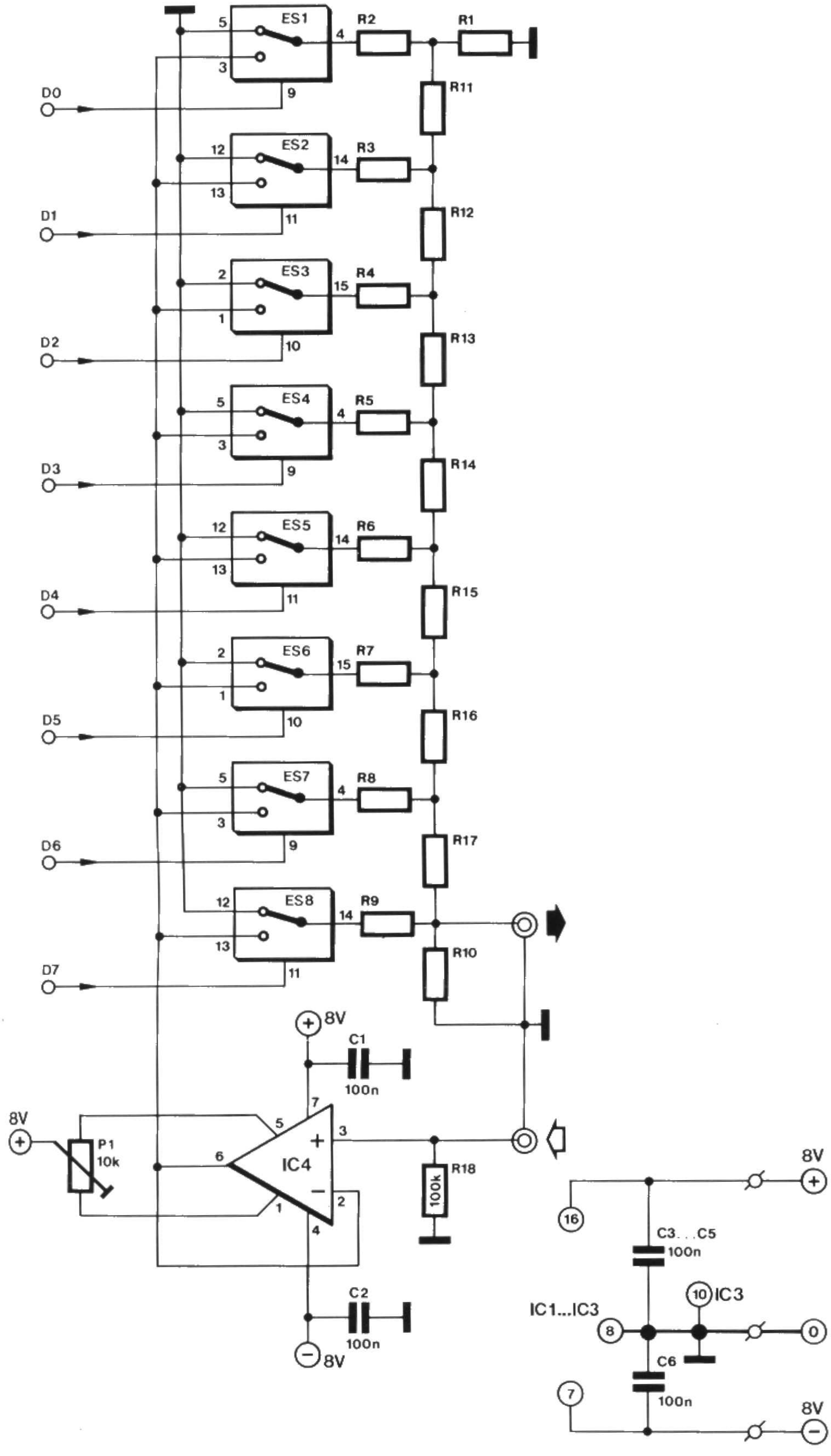
Currently, 48 MHz quartz crystals are widely available at relatively low cost thanks to their use in computer systems. In these, there is often a need for several clock frequencies that can be derived from a central oscillator. When this supplies a buffered 48 MHz signal, it is relatively simple to add a divider circuit that provides lower, phase-synchronous, clock signals of, say, 6, 8, 12, 16 or 24 MHz. Obviously, this obviates the need for separate quartz crystals and associated oscillators, and so economizes on hardware expenses.

A reliable 48 MHz oscillator is fairly difficult to make with HC or HCT gates. The oscillator shown here is, therefore, built around discrete RF transistors. It operates with inexpensive, third overtone series resonance quartz crystals in the range between 44 MHz and about 52 MHz. A parallel L-C network may be



connected in series with the crystal as shown in the circuit diagram for accurate setting of the oscillation frequency to 48.000 MHz, but also for 'pulling' the oscillator a few kilohertz around this frequency if it is used for driving a frequency multiplier (local oscillator chains in 2 m, 70 cm, or 23 cm amateur radio equipment).

056 DIGITAL ATTENUATOR



IC4 = LF 356
ES1...ES3 = IC1
ES4...ES6 = IC2 } 4053
ES7, ES8 = IC3
R1...R10 = 22k, 1%
R11...R17 = 11k, 1%

884083-10

Digitally controlled attenuators almost invariably use some kind of tapped resistor network to simulate a potentiometer. This solution is fine as long as the number of steps required is small. When finer control is required, however, the normal tapped resistor network is hardly ever used because of the large number of components that would be required. The circuit shown here offers relatively high resolution (attenuation range: 48 dB) whilst requiring few components only.

The technique used is similar to that of multiplying DACs (digital-to-analogue converters). In a conventional R-2R ladder DAC, the output voltage is given by $(U_{ref}/384)N$, where N is the binary number applied to the inputs. The direct dependance of the output voltage on U_{ref} makes it easy to obtain a variable attenuator by substituting the input for U_{ref}. The output will then be $(U_{in}/384)N$.

The R-2R ladder network used here is composed of resistors R_1 to R_{17} incl., while electronic switches ES_1 to ES_8 incl. form the switching elements. These are of the two-way type (SPDT), connecting either the input voltage or ground to the inputs of the ladder network. Buffer IC_1 presents a constant impedance to the source. Pin 7 of IC_1, IC_2 and IC_3 should be grounded unless the circuit is operated with bipolar signals. In that case, pin 7 of all three ICs is connected to −8 V.

The circuit can handle signals of up to 400 kHz with a maximum amplitude of about 4 V_{pp}. With signals of lower level, higher frequency response should be obtainable. The high frequency limit is due to the buffer at the input—the electronic switches by themselves can handle signals up to 10 MHz.

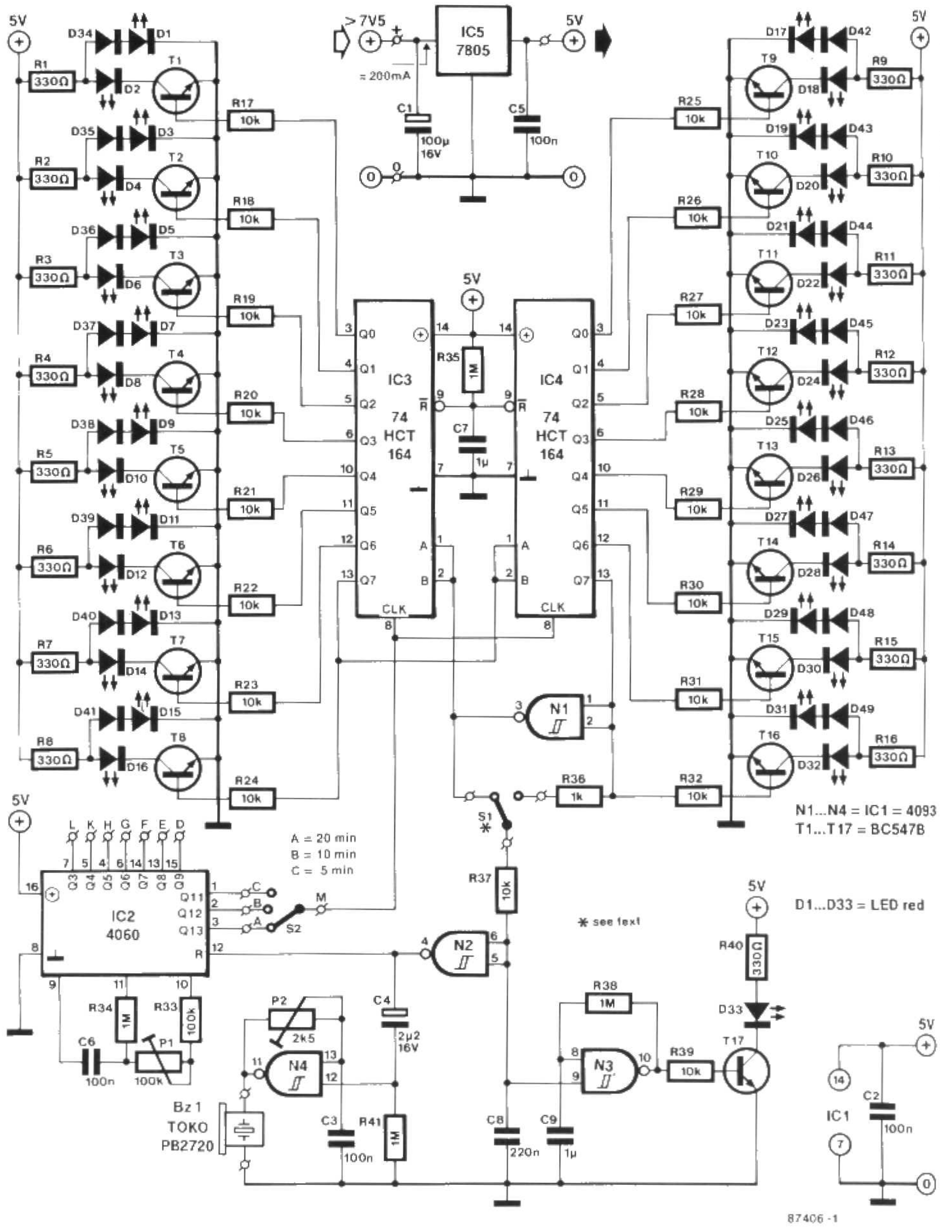
The fixed attenuation of the circuit is about −3.5 dB. Signal-to-noise ratio is more than 100 dB at an input signal of 1 V_{rms}. The output offset voltage is compensated by adjusting P_1. Current consumption of the circuit is about 6 mA at $U_b = \pm 8$ V. Finally, it should be noted that TTL circuits can not drive the circuit direct, unless 47 kΩ pull-up resistors are fitted at control inputs DØ to D7 incl.

057 ELECTRONIC SAND-GLASS

This electronic version of the reversible sand-glass uses a set of LEDs to simulate the passing of sand grains from the upper to the lower bulb. The simple to build circuit is accurate enough for most domestic timing applications.

The circuit diagram appears in Fig. 1. On power-up, shift registers IC_3 and IC_4 are reset by the low pulse from network R_{35}-C_7. A few seconds later, the sand-glass is started. The oscillator in IC_2 generates a clock signal for the shift registers. The clock frequency is adjustable with P_1. Switch S_2 enables selecting one of the three timing periods stated in the circuit diagram. S_1 is a small mercury or ball changeover switch mounted inside the sand-glass. When this is reversed, the switch toggles and so selects the odd or even numbered LEDs. Assuming that S_1 is set as shown in the circuit diagram, every clock pulse causes a logic high level to be shifted into IC_3, for as long as pin 13 of IC_4 remains logic low. The MS bit of IC_3 (output Q7) is shifted into the second shift register, IC_4. Controlled by the shift register outputs, transistors T_1 to T_{16} incl. switch off the odd numbered LEDs, and light the even numbered ones

1



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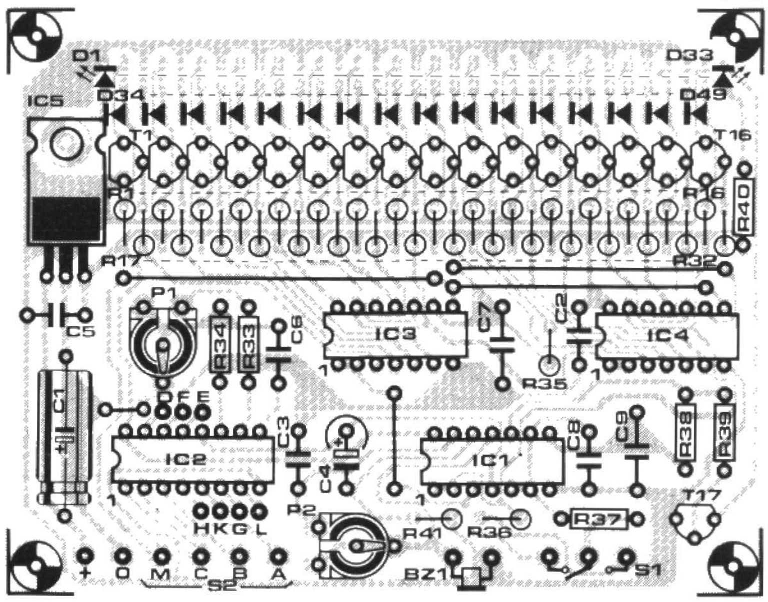


Parts list

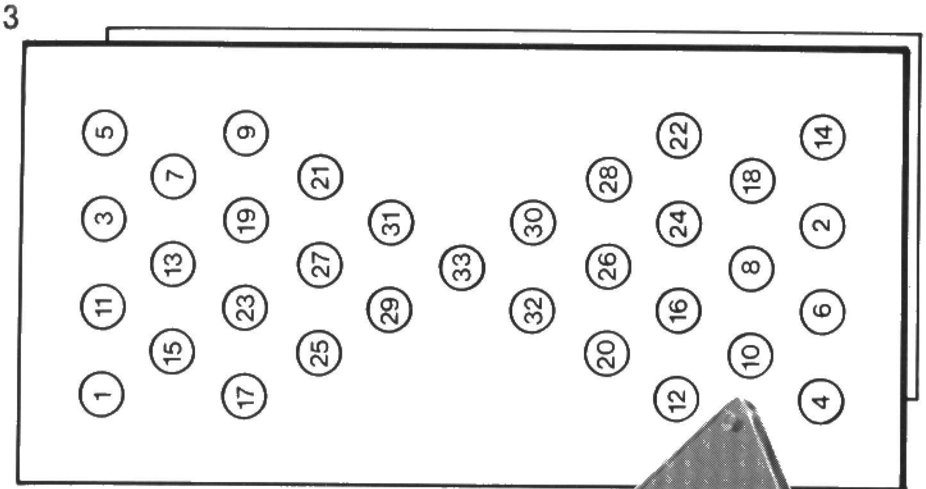
Resistors (±5%):
R_1 . . . R_{16} incl.; R_{40} = 330R
R_{17} . . . R_{32} incl.; R_{37}; R_{39} = 10K
R_{33} = 100K
R_{34}; R_{35}; R_{38}; R_{41} = 1M0
R_{36} = 1K0
P_1 = 100K preset
P_2 = 2K5 or 2K2 preset

Capacitors:
C_1 = 100μ; 16 V; axial
C_2; C_3; C_5; C_6 = 100n
C_4 = 2μ2; 16 V; radial
C_7; C_9 = 1μ
C_8 = 220n

Semiconductors:
D_1 . . . D_{33} incl. = red LED
D_{34} . . . D_{49} incl. = 1N4148
T_1 . . . T_{17} incl. = BC547
IC_1 = 4093
IC_2 = 4060
IC_3; IC_4 = 74HCT164
IC_5 = 7805

Miscellaneous:
Bz_1 = PB2720 (Toko; Cirkit stock no. 43-27201).
S_1 = SPDT mercury, ball or tilt switch, e.g. Maplin order no. FE11M, or ElectroValue no. 339-881.
S_2 = single-pole, 3-position rotary switch plus knob.
PCB Type 87406 (available through the Readers Services).
Suitable ABS enclosure.
DC power socket.

3



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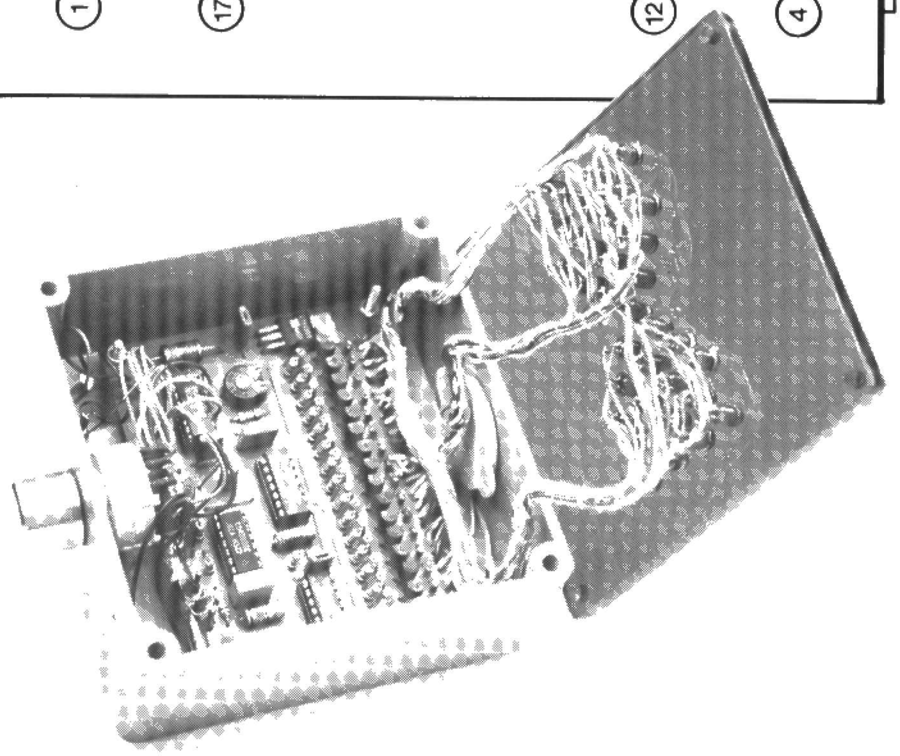
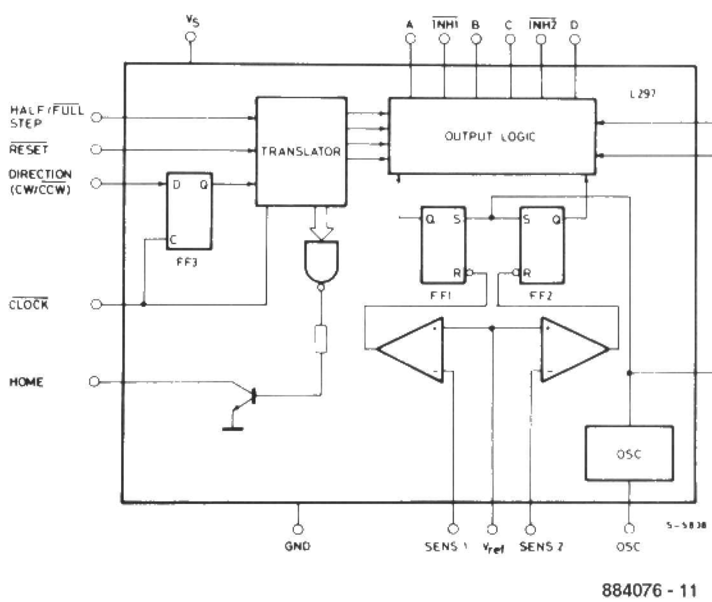


sequentially. When pin 13 of IC_4 goes high, counter IC_2 is reset via N_1-N_2, while oscillator N_4 is started. Buzzer Bz_1 is actuated and sounds for about 2 seconds (C_4-R_{41}). The pitch of the tone can be set with P_2.
When the sand-glass is reversed, S_1 toggles, ending the reset state of IC_2. Logic low levels are shifted into IC_3 because pin 13 of IC_4 is logic high. The even numbered LEDs go out one by one, and the odd numbered ones light, until pin 13 of IC_4 goes low again. IC_2 is reset, Bz_1 produces a short beep, and the sand-glass can be reversed for a new timing period. LED D_{33} indicates that the sand-glass is operative. The circuit is fed from a small mains adaptor capable of supplying about 200 mA at an output voltage between 7.5 and 12 VDC.
Construction of the sand-glass is straight-forward using PCB Type 87406—see Fig. 2. The position of the LEDs on the front panel of the enclosure is shown in Fig. 3. Make sure that each LED is connected to the corresponding soldering island on the PCB. SPDT Switch S_1 is made from two SPST mercury or ball switches, fitted together but mutually reversed at a suitable position in the enclosure. The action of the switches is tested by reversing the sand-glass and measuring the switch configuration with the aid of a continuity tester or an ohm meter. All parts in the sand-glass enclosure should be fitted securely in view of the reversibility of the enclosure. The socket for connecting the adaptor, and rotary switch S_2, are fitted in one of the side panels. A prototype of the electronic sand-glass is shown in Fig. 4. The detachable front panel that holds the LEDs was cut from perspex sheet.

058 STEPPER MOTOR DRIVER

This stepper motor driver is simple without compromising versatility. It has a conventional computer interface, an optional on-board clock generator, and integrated motor control and driver stages. The board can drive dual-stator motors, i.e., motors with two bipolar, or four unipolar, phases. Maximum current per phase is 2A.
The circuit diagram of Fig. 2 shows that the circuit is designed around chip set L297-L298 from SGS.
The block diagram of the L297 is given in Fig. 1. This IC generates control signals for a dual-stator motor, and enables selecting direction of travel and full or half-step operation by appropriate programming of its TLL-compatible inputs. One full or half step is performed on the trailing edge of the signal applied to the clock input (CLK). When the enable input, E, is held logic low, the motor is not energized, so that the spindle can be rotated freely. Driv-

1



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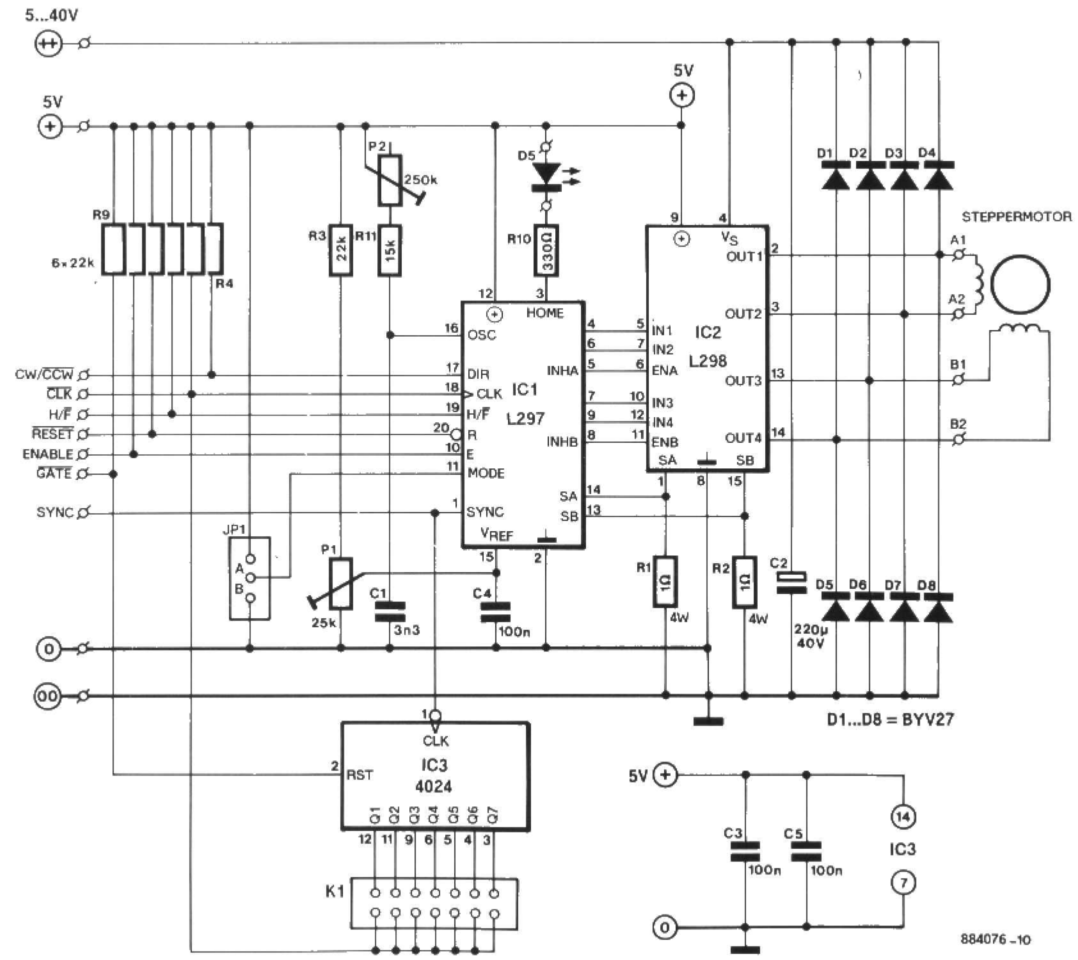
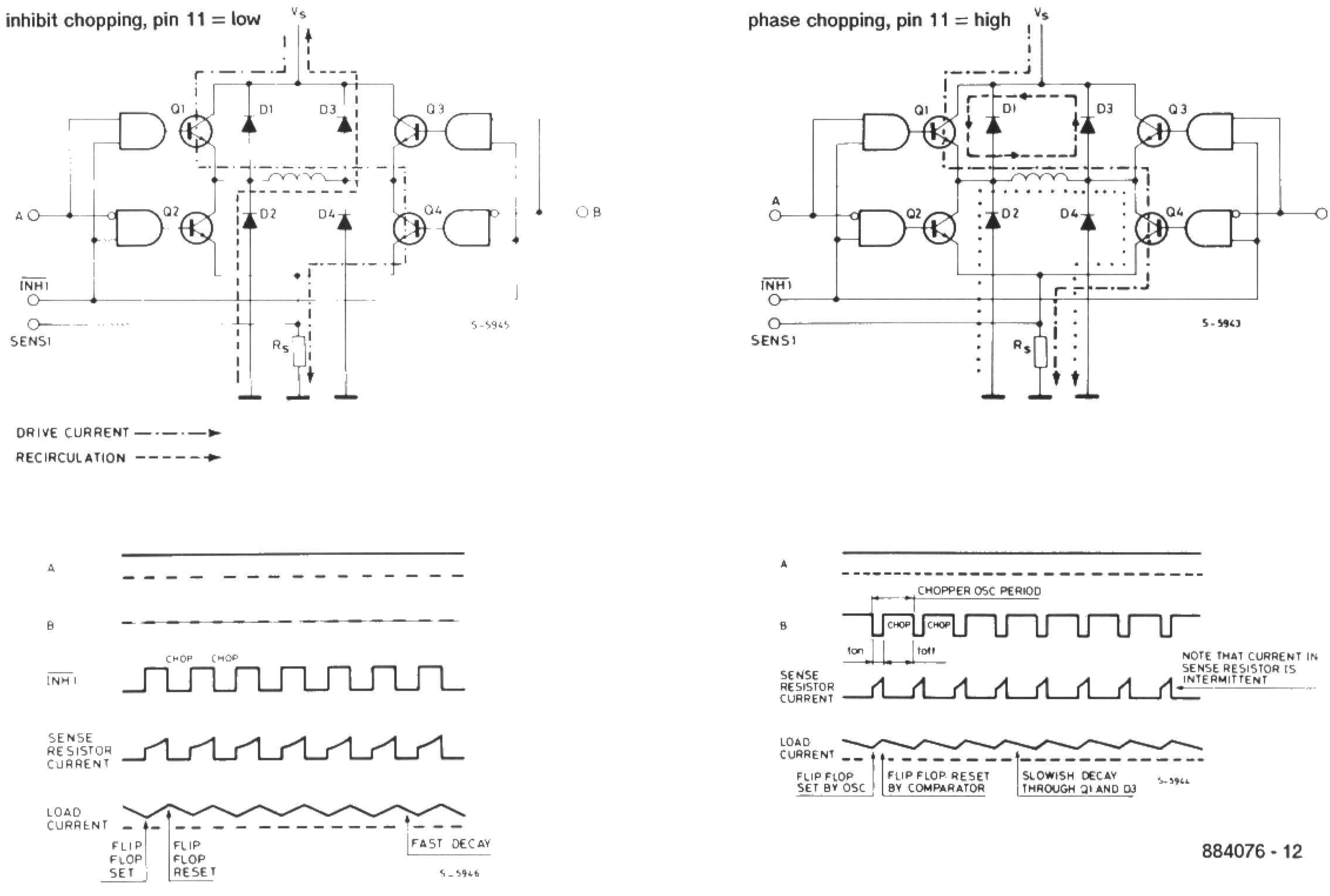


5...40V
5V
R9
6×22k
R4
R3 22k
R11 15k
P2 250k
D5
R10 330Ω
CW/CCW
CLK
H/F
RESET
ENABLE
GATE
SYNC
JP1
A
B
P1 25k
C1 3n3
C4 100n
IC1 L297
OSC DIR CLK H/F R E MODE SYNC VREF HOME INHA INHB SA SB
IC2 L298
IN1 IN2 ENA IN3 IN4 ENB SA SB VS OUT1 OUT2 OUT3 OUT4
D1 D2 D3 D4
D5 D6 D7 D8
STEPPERMOTOR
A1 A2 B1 B2
R1 1Ω 4W
R2 1Ω 4W
C2 220µ 40V
D1...D8 = BYV27
IC3 4024
CLK RST Q1 Q2 Q3 Q4 Q5 Q6 Q7
K1
C3 100n
C5 100n
IC3
884076-10

3



inhibit chopping, pin 11 = low
phase chopping, pin 11 = high
Vs
Q1 Q2 Q3 Q4
D1 D2 D3 D4
A
B
INH1
SENS1
Rs
5-5945
5-5943
DRIVE CURRENT
RECIRCULATION
A
B
INH1
CHOP CHOP
SENSE RESISTOR CURRENT
LOAD CURRENT
FLIP FLOP SET
FLIP FLOP RESET
FAST DECAY
5-5946
CHOPPER OSC PERIOD
CHOP
ton
toff
NOTE THAT CURRENT IN SENSE RESISTOR IS INTERMITTENT
FLIP FLOP SET BY OSC
FLIP FLOP RESET BY COMPARATOR
SLOWISH DECAY THROUGH Q1 AND D3
5-5944

884076 - 12

Parts list

Resistors (±5%):
R_1;R_2 = 1R0; 4 W
R_3 . . . R_9 incl. = 22K
R_{10} = 330R
R_{11} = 15K
P_1 = 25K or 22K preset H
P_2 = 250K or 220K preset H

Capacitors:
C_1 = 3n3
C_2 = 220μ; 40 V
C_3;C_4;C_5 = 100n

Semiconductors:
D_1 . . . D_8 incl. = BYV27 (Philips Components)
D_9 = LED
IC_1 = L297 (SGS)
IC_2 = L298 (SGS)
IC_3 = 4024

Miscellaneous:
JP_1 = 3-way terminal strip (pitch: 0.1 in.); 1 jumper.
K_1 = 2 off 7-pin terminal strips (pitch: 0.1 in.); 1 jumper.
14 off solder pins dia. 1.3 mm.
PCB Type 884076 (see Readers Services page).

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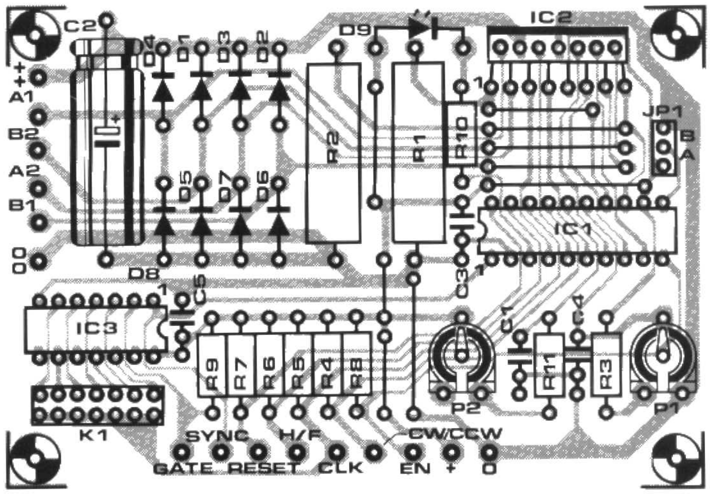

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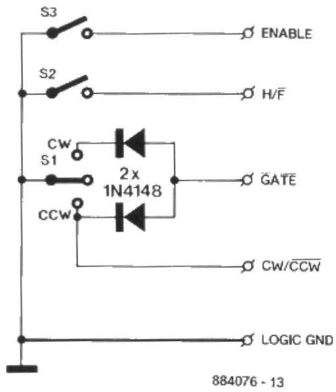

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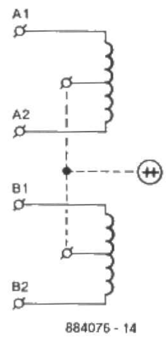

ing the reset input logic low causes the motor to remain halted in the home position (LED D_9 is quenched).
Power driver Type L298 supports constant current drive of the stator windings. Current drive gives good results because it allows stepper motors to be connected to a voltage that is higher than specified for voltage drive. Current drive considerably improves the motor's dynamic characteristics (start frequency and maximum step-rate). An internal oscillator sets a bistable at the start of each period, when the stator windings are connected to the supply voltage. Due to the stator inductance, output current will initially rise linearly, resulting in a linear voltage on current sensing resistors R_1 and R_2. When the measured voltage reaches a certain user-defined peak value, V_{ref}, two internal comparators reset the bistables, and the stator current is interrupted. Free-wheeling diodes then reduce the induced stator field. From the above it is clear that current drive works by peak detection. The resultant avarage current depends on V_{ref} (adjustable with P_1), the oscillator frequency (adjustable with P_2) and the values of the sensing resistors. Ripple amplitude on the stator current depends on stator self-inductance and the logic level at the MODE input. When this is high, the outputs of IC_2 are switched to high impedance during the free-wheeling period. The stator field is reduced fairly rapidly via the free-wheeling diodes which conduct because the instantaneous voltage on the stator winding is slightly higher than the supply voltage. When MODE is held logic low, one transistor in the bridge circuit internal to the L298 remains on during the free-wheeling period. This causes the free-wheeling voltage on the stator winding to remain relatively low, resulting in slower reduction of the stator field strength and, therefore, reduced ripple (*phase chopping*, see Fig. 3). This option is offered to enable efficient current control of motors with a relatively low stator self-inductance.
Synchronization of the oscillators in the L297s is required when multiple drivers and motors are used in a single system. This is simple to accomplish by fitting parts P_2, R_{11} and C_1 on one driver board only, and feeding the signal available at the SYNC output to the SYNC terminal on the other boards.
An on-board divider, IC_3, is provided to supply the clock signal when the relevant computer output line cannot be programmed to toggle at the required step-rate. The divider is clocked with the SYNC signal of the L297, and jumper block K_1 allows selecting 1 of 7 available clock frequencies (step-rates). On-board clocking via IC_3 can be disabled by driving input GATE logic low. The CLOCK input then functions as an output, enabling the computer to keep track of the number of steps performed. When external clock pulses are applied to the board, IC_3 is simply omitted.
The 5–40 V supply rail need not be regulated — smoothing is adequate here. The maximum attainable step-rate increase with supply voltage, but 40 V should not be exceeded.
The chopper frequency (refer to Fig. 3), and hence the step-rate in stand-alone applications, is set with P_2. Stator current is set with P_1. Lisping sounds produced by the motor point to instability of the current drive. This effect can be remedied by either re-adjusting the chopper frequency, or by selecting the other logic level at the MODE input of IC_1. When this still fails to stabilize the current drive, the supply voltage must be reduced until the motor operates with voltage instead of current drive.
Stand-alone use of the driver is simple to accomplish by connnecting three external switches as shown in Fig. 5. Figure 6 shows how to connect the driver board to a unipolar motor. The oscillator inside IC_1 is used only for generating the clock signal required in stand-alone applications of the driver. When it is used, the step-rate can be set by fitting a jumper in the appropriate position on K_1, and adjusting P_2.
Finally, IC_2 is purposely located at the edge of the printed circuit board to enable it to be bolted on a metal surface for cooling.

059

VIDEO DISTRIBUTION AMPLIFIER

The Type TEA5114 from Thomson-CSF comprises three electronic switches followed by a buffer/amplifier. Normally the voltage amplification is 2 (6 dB). When the input voltage exceeds 1.2 V_{pp}, or when the output voltage exceeds 1.5 V_{pp}, an internal selector reduces the amplification to unity (0 dB). The threshold of 1.2 V_{pp} is created with the aid of voltage divider R_4-R_5, which also forms the input termination of 75 Ω. Series resistors R_1-R_3 ensure 75 Ω output impedance for driving video equipment via standard coax cable. The TEA5114 can be used as a video source selector also, provided each input has its own 75 Ω termination network. The non-connected inputs should then be fitted with a coupling capacitor. Channel selection is effected by controlling the logic level at pins 10, 12 and 15. Note that the logic 1 (high) level corresponds to +2.5 V here.

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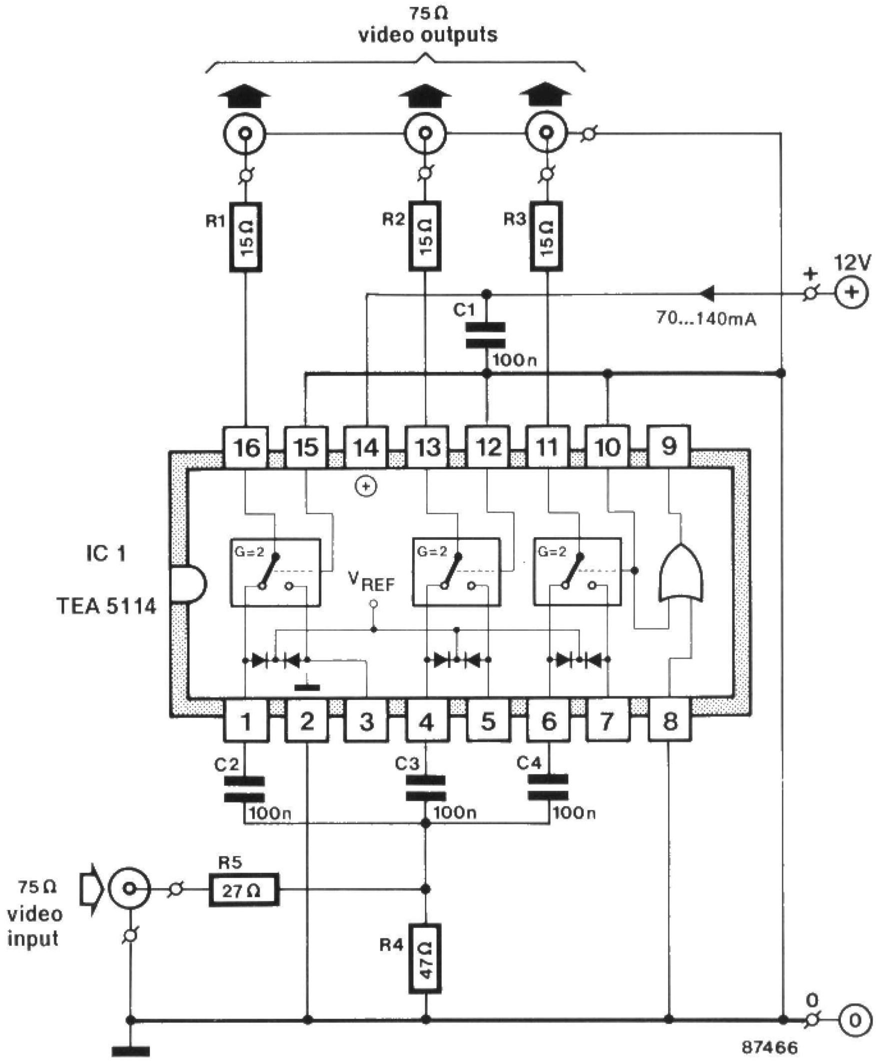

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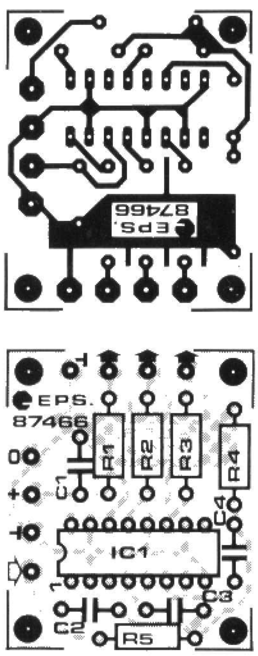
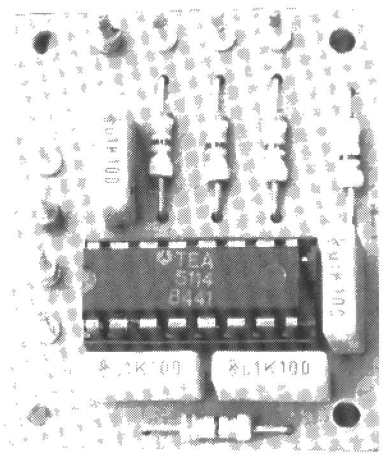




Parts list

Resistors (±5%):
R_1 . . . R_3 incl. = 15R
R_4 = 47R
R_5 = 27R

Capacitors:
C_1 . . . C_4 incl. = 100n

Semiconductor:
IC_1 = TEA5114*

Miscellaneous:
PCB Type 87466 (not available through the Readers Services).

* Thomson Components Limited • Ringway House Bell Road • Danneshill • Basingstoke • Hants RG24 0QG. Telephone: (0256) 29155. For distributors see Infocard 502 (*EE* February 1987).

060 DISCRETE +5 to −15 V CONVERTER

This negative voltage converter differs from a host of other designs in not being set up around the latest integrated circuit. The circuit diagram shows that only a handful of commonly available parts are required to build an efficient +5 to −15 V converter.

IC_1 functions as a self-oscillating multivibrator that supplies an output signal with a relatively high duty factor. The LM311 is designed to operate from a single 5 V supply, and has a high output current capability for driving switching transistor T_1. Duty factor of the output signal is determined mainly by voltage divider R_2-R_3, and frequency of oscillation by C_2-R_4. Transistor T_2 forms part of a regulation loop that modifies the oscillator duty factor to maintain −15 V at the output of the converter.

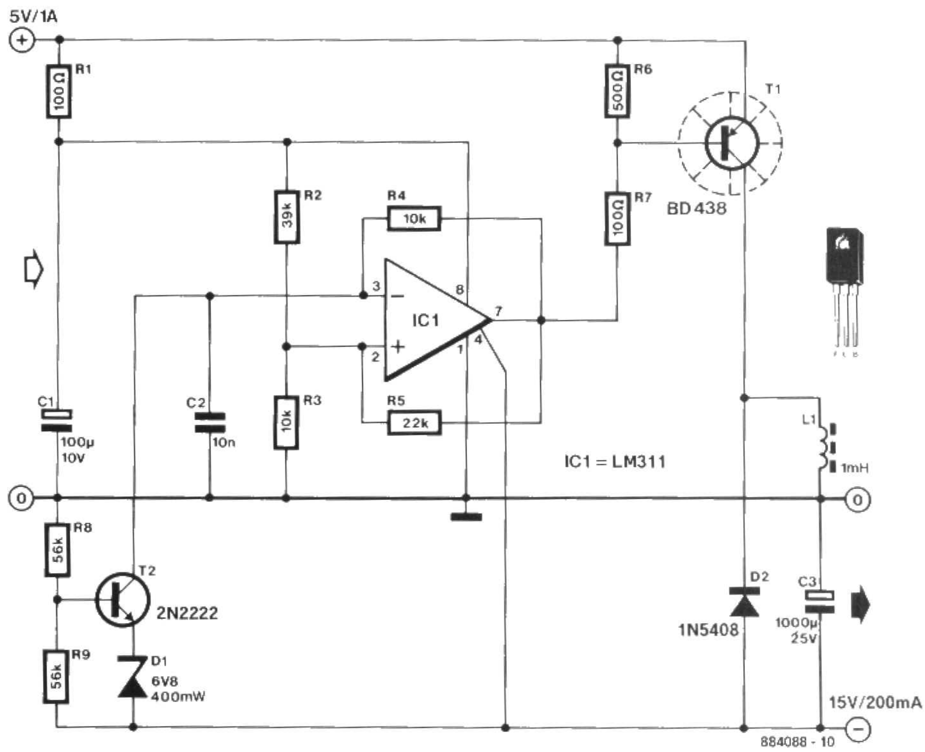
The output voltage, U_o, is calculated from

$$U_o = -(U_{D1} + U_{B\text{-}E(T1)})(R_8/R_9 + 1) \qquad [V]$$

The component values shown give the following design data:

Efficiency (P_o/P_i):	max. 75%
Oscillator frequency:	6 kHz
Duty factor:	approx. 0.8
Output ripple voltage:	100 mV at I_L=200 mA
Maximum load current:	200 mA

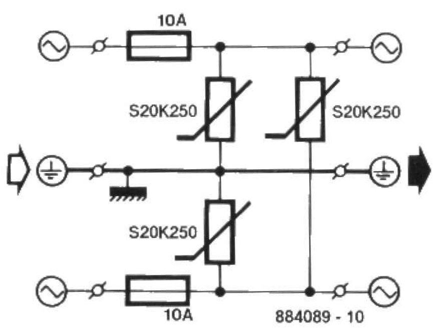
T_1 should be fitted with a small heat-sink.



Source: *National Semiconductor Linear Brief 18.*

061 OVER-VOLTAGE PROTECTION

It happens from time to time that very large voltage spikes (lightning; switching of large loads) are superimposed on the mains. Although these spikes are of very short duration, they may have disastrous consequences for mains-operated equipment. A mains power supply can be effectively protected from such spikes with the aid of varistors. These components can handle, but only for a few microseconds, currents of thousands of amperes. In the proposed protection circuit, three varistors are used: one between L(ive) and N(eutral); one between L en E(arth); and one between N and E. The varistors are preceded by fuses, so that only the equipment connected via the circuit is protected. If these fuses were omitted, the entire household supply would be protected with the risk that one of the main fuses blows during an over-voltage.



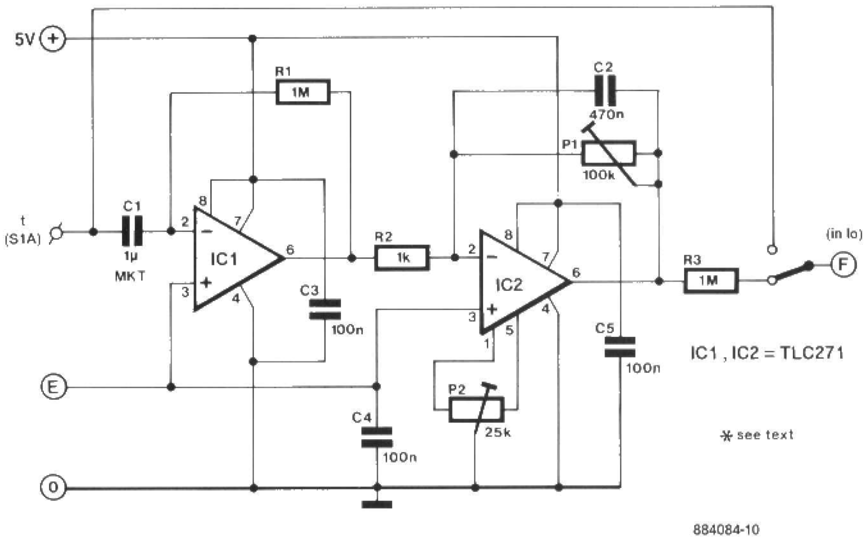
The circuit is best built into a small man-made-fibre enclosure with integral plug and socket. The mains-carrying bare wires should be kept separated by at least 3 mm.

062 FROM ALTIMETER TO VARIOMETER

The altimeter published some 18 months ago [1] can be adapted to function as a variometer by the following circuit. The difficulty in the design of the circuit is, of course, that it has to work with very small input voltages. It is based on the fact that differentiating the absolute height gives as result the rate of change of altitude.

In the diagram, IC_1 is the differentiator that operates with a time constant, R_1C_1, of 1 s. Since this type of differentiator inverts, it is followed by an inverter.

If the amplification is arranged at 60 (P_1 set to 60 k), the eventual read-out shows the rate of change of altitude in m/min, assuming, of course, that the altimeter has been calibrated as prescribed in Ref. 1.



Because of the very low levels of signal input, the choice of components is critical. For instance, C_1 must be an MKT, not an electrolytic, type. The differentiator is a CMOS opamp that not only has a very high input impedance, but also extremely small drift of offset voltage with temperature. This drift is so small only if the opamp is used in the low-bias mode (pin 8 connected to +). This has the additional benefit of very low current (typically 10 μA). It also has a disadvantage in that the slew rate is only 0.04 V/μs, but that does not matter here, since for all practical purposes the stage functions as a d.c. amplifier.

Offset voltages are also undesirable in IC_2, because they are added to those of IC_1 and appear amplified at the output. Therefore, P_2 has been incorporated to compensate all offset voltages. Preferably, IC_2 should also operate in the low-bias mode, but it may be found necessary to connect pin 8 to pin 3 (medium bias) or even pin 4 (high bias) to obtain full offset compensation. This has to be tried out in each and every individual unit. Simply adjust P_2 for a display reading of 000 when the unit is at rest.

The terminals should be connected to the corresponding ones in the altimeter. The switch at terminal F allows selection between altimeter and variometer use.

The add-on circuit may conveniently be mounted above or under the altimeter PCB, and should be well screened.

When the unit is used as variometer, the multiturn potentiometer (P_7) in the altimeter must not be turned. If the unit is used as barometer, the switch should be set to the altimeter position.

Readers should note that the circuit has been tested in laboratorium conditions only and NOT in practical use.

Reference: *Elektor Electronics*, November 1986, pp. 20—26.

063 BACKGROUND-NOISE SUPPRESSOR

Hiss, crackling, and other discordant sounds are disconcerting and frequent sources of annoyance to most music lovers. Unfortunately, the sources of this background noise are not easy to eliminate, but the circuit proposed here will be of help. It should be appreciated, however, that the suppression of noise is always a last resort: the best way of getting rid of it at source.

The circuit is based on the fact that background noise is always at its most annoying during quiet music passages. It attenuates the output signal by some 45 dB when there is no or very low music signal input. When the input rises, the attenuation decreases proportionally, becoming 0 dB with normal to loud passages.

The input signal is taken direct to the output terminals via R_{11} and R_{12} respectively. At the same time, they are summed via R_1 and R_2 and applied to non-inverting amplifier IC_1 via potentiometer P_1. The cross-over point in the gain characteristic of IC_1 is determined by R_5 and C_1. Frequencies above the cross-over point are not amplified, and so do not contribute to the suppression. The output of IC_1 is rectified by D_1—D_2 and used to switch off T_1. This enables T_2 and T_3 to short-circuit the output and thus suppress the noise signal.

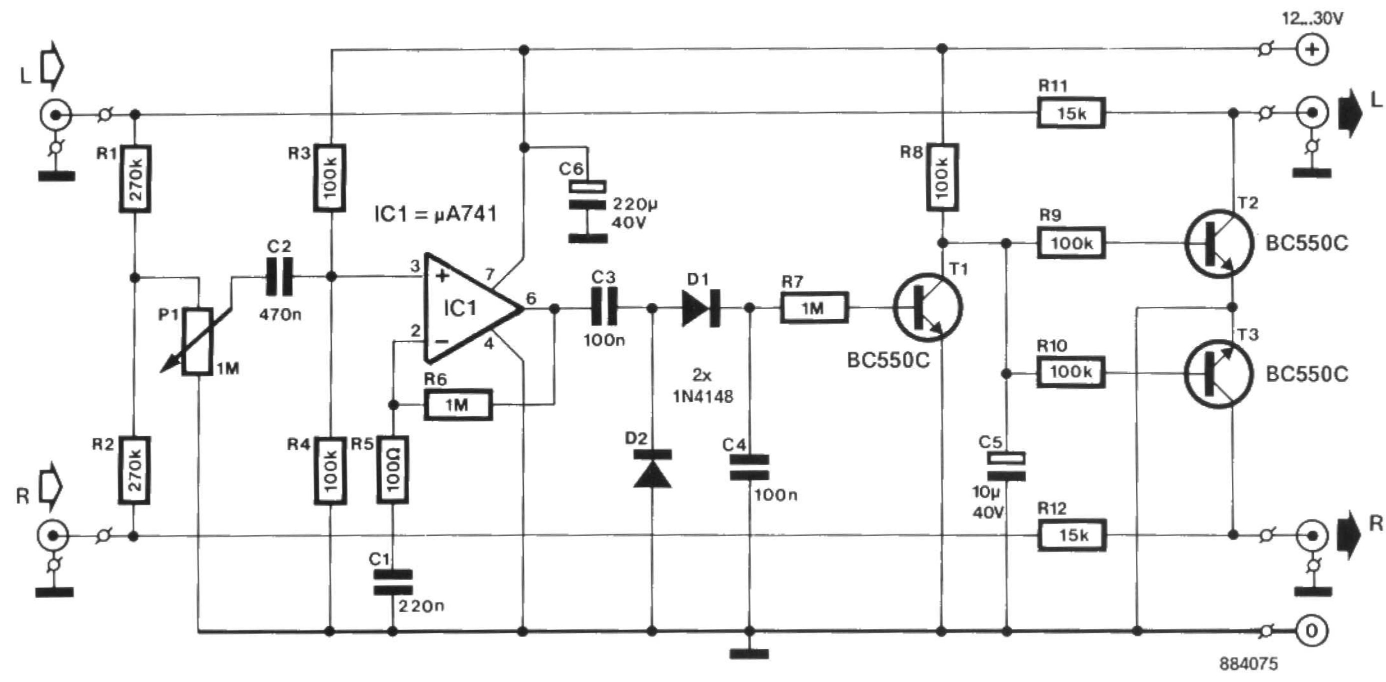
When T_1 begins to conduct, the base voltage of both T_2 and T_3 decreases and the output attenuation is reduced: noise signals are thus suppressed to a lesser degree.

The sensitivity of the circuit may be varied by P_1: the higher the sensitivity, the sooner the suppression lessens. This allows the sensitivity to be matched to different music sources.

The peak signal level the circuit can handle is about 210 mV. Distortion at that level is not greater than 0.01%.

The delay before the circuit operates is determined by time constant R_7C_4. With values as shown, it is about 1 s but can, of course, be altered to individual taste.

The circuit operates from a 12—30 V supply and draws a current of 2 to 3 mA.



064 D.C. DETECTOR

The d.c. component of a signal can only be detected by separating it from the a.c. component. This is most conveniently done by filtering the a.c. component. In the proposed circuit this is effected with the aid of the common-mode rejection ratio (CMRR) of an opamp. (The CMRR is a measure of the ability of the opamp to produce a zero output for like inputs).

The complete signal is applied to the inverting input of opamp A_1, and only the a.c. component, via C_1, to the non-inverting input. The lowest frequency that can be detected is determined by time constant $(R_3+R_4)C_1$. With values as shown, a.c. suppression amounts to about 50 dB at 20 Hz.

The output of A_1 is fed to a low-pass filter to further attenuate high frequencies. This is necessary because the CMRR of an opamp decreases at higher frequencies. The difference signal is then applied to comparator A_2. Diodes D_1 and D_2 ensure that A_2 reacts only to voltages greater than ± 300 mV.

A negative direct voltage at the input of the circuit results in a positive potential at the inverting input of A_2, which causes the relay to be deactuated (it is normally energized as long as the 12 V supply is on). A positive direct voltage at the input results in a negative potential at the non-inverting input of A_2, so that, again, the relay is deactuated.

In normal operation, the voltage at the non-inverting input of A_2 is arranged by potential divider R_7-P_1-R_8-R_9, so that the relay is energized. Because of C_8, the relay is energized a few seconds after the supply has been switched on.

Capacitors C_3 and C_4 serve to smooth low-frequency signals so as to prevent clattering of the relay.

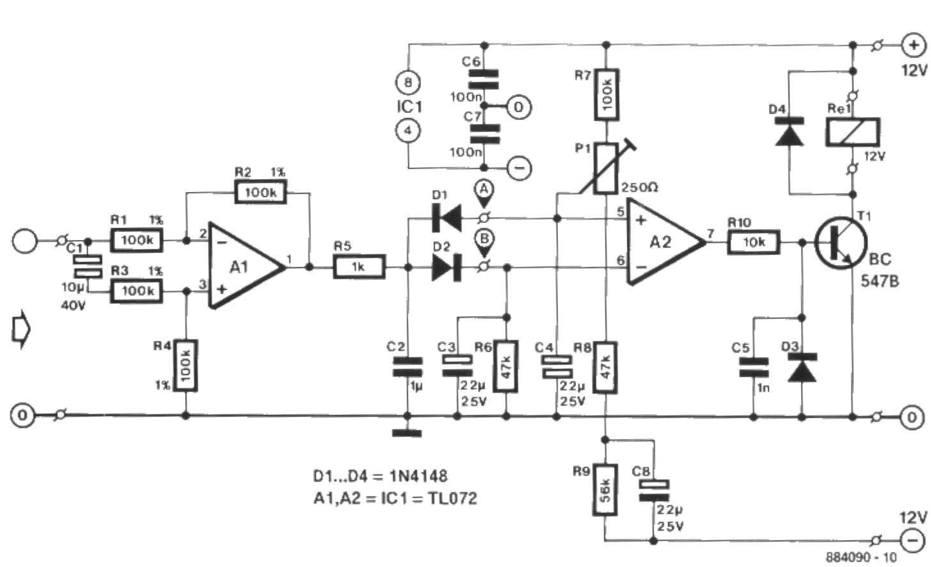
The relay is driven by a BC547B which can switch currents up to 100 mA. The supply to the relay should be not higher than 18 V.

If the circuit is powered by a not entirely symmetrical supply, it may happen that the travel of P_1 is insufficient: the value of R_7 should then be altered as required.

When the circuit is used in an active loudspeaker system, each output stage should have its own detector, consisting of A_1 and associated components up to points A and B in the diagram. The outputs of these detectors are then connected in parallel to A and B.

For mid- and high-frequency sections of the loudspeaker system the time constant of the input to A_1 may be made smaller to obtain a faster reaction to d.c. components.

Finally, the current drawn by the circuit is determined primarily by the relay.



065 STEPPED VOLUME CONTROL

The circuit consists of three distinct sections. The first consists of a straightforward amplifier, IC_{1a} and IC_{1b}. The second is a digital counter, IC_3, which converts a binary code into a resistance value via IC_2. That value is used to control the degree of amplification. Finally, there is a pulse shaper, IC_4, which enables IC_3 to count up or down.

Amplifier IC_{1b} has a switch-controlled gain of 0 dB or 24 dB. The control switch, S_3, is an electronic type driven from output Q_d of IC_3.

The gain of IC_{1a} can be set between 0 dB and 21 dB in steps of 3 dB. The total gain of the two amplifiers can thus be set between 0 dB and 45 dB.

The bandwidth of the amplifier extends from 10 Hz to 40 kHz. The peak value of the amplified signal should not rise above 8 V_{pp} with a supply voltage of 5 V.

The pulse shaper, formed by bistable N_1-N_2 and network C_5-R_{16}, indicates to IC_3 whether it should count up or down. The RC networks suppress spurious pulses.

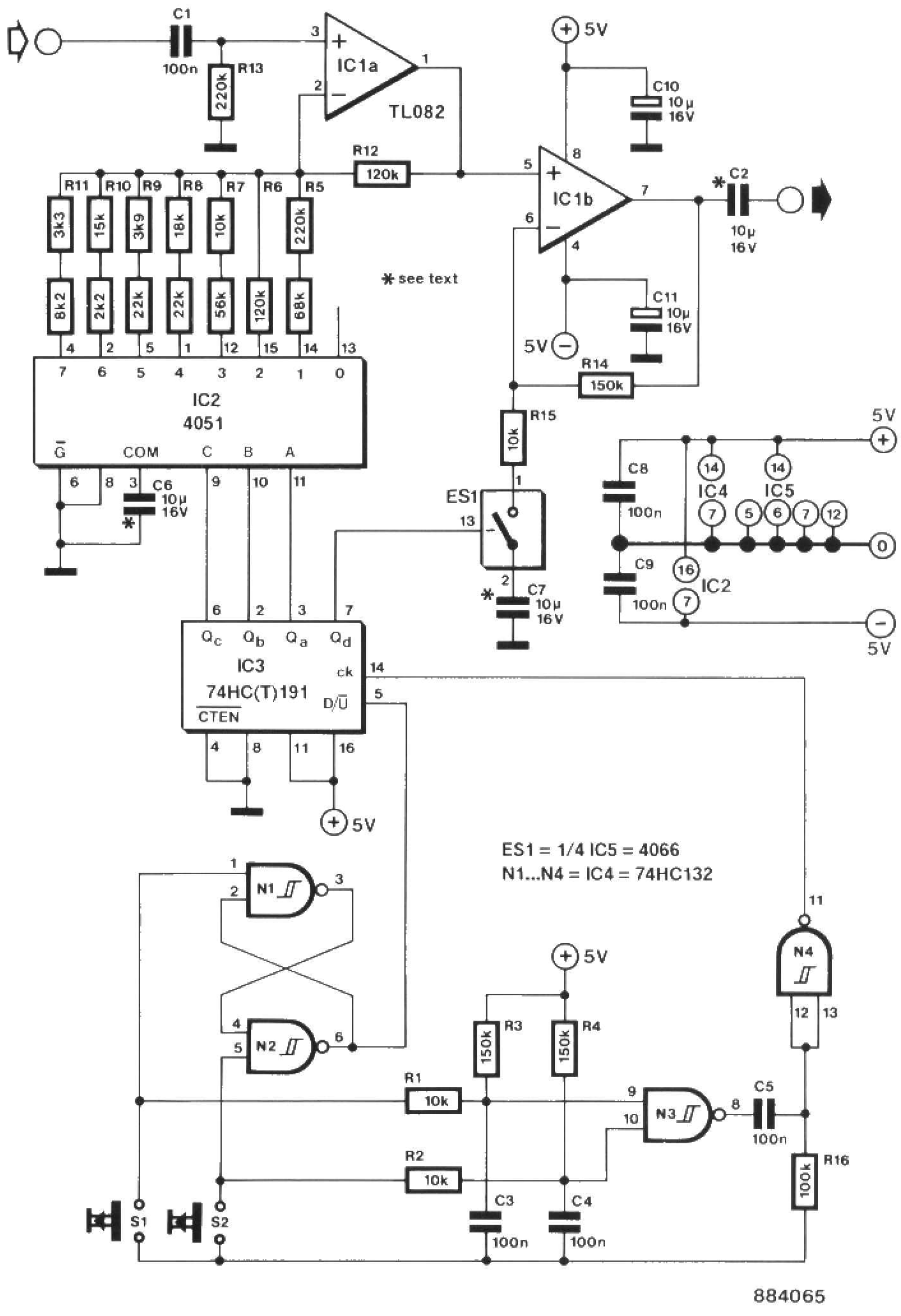
The delay introduced by the RC networks before N_3 ensures that the clock pulse can not appear at the clock input of IC_3 before the direction of counting has been set. The count position can, therefore, be increased or reduced by switches S_1 or S_2 respectively.

Inputs C_k and $D/\overline{U}$ of IC_3 may be used to connect a software potentiometer: a two-wire connection per control is sufficient. An 8-bit user port can thus accommodate four of such digital potentiometers.

A standard CD4051 must be used for IC_2, because HC or HCT types do not allow the use of a negative supply voltage on pin 7. The other ICs may be HC or HCT types. If an LS type is used for IC_3, 4k7 pull-up resistors are necessary at the outputs of this circuit to match the voltage levels of the two logic families.

Note that C_2, C_6, and C_7 are bipolar electrolytic types.

The total current drawn by the circuit is about 10 mA.



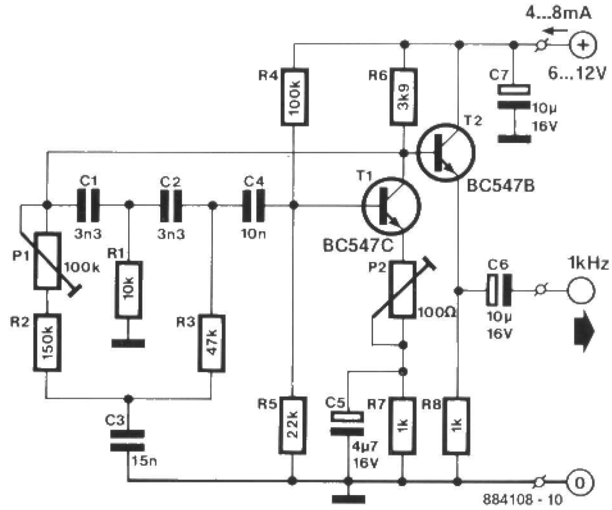
066 NOSTALGIC SINE WAVE GENERATOR

As far as young engineers and technicians are concerned, a sine wave generator is something you make from an XR2206. In the pre-IC era, sine wave generators were designed around discrete components. The generator described here has, however, more than just nostalgic value: it is also educational (and perhaps suitable for writers of the history of electronics).

The (fixed) output frequency is fairly stable at 1 kHz, and the distortion, after proper adjustment, below 1%. The generator is suitable for use as an audio test generator or as a morse code trainer and costs only a couple of pounds to make.

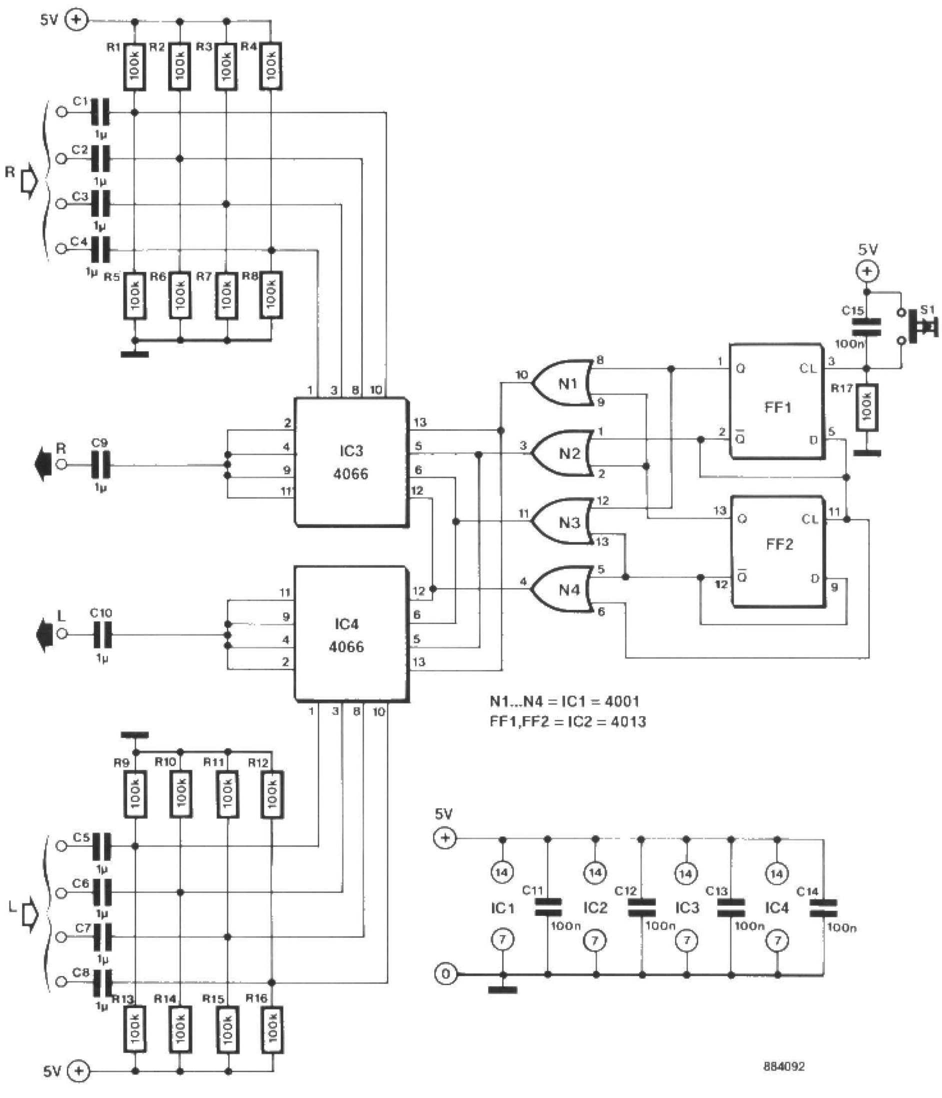
The generator is of the so-called

double-T type, which has the advantage of not needing any inductors. The oscillator proper, T_1, is followed by an emitterfollower, T_2, which ensures a sufficiently low output impedance.
The frequency is set to 1 kHz by P_1, and P_2 minimizes the distortion of the waveform. With P_2 set for minimum resistance, the amplitude of the output signal will be maximum, but the distortion will be quite appreciable. Increasing the resistance will reduce the distortion, but it may happen that when P_2 is nearing its maximum value oscillations stop. Setting P_2 is thus finding a compromise between acceptable distortion and reliable oscillations. The output level also depends on the setting of P_2: it lies between 1.5 V_{pp} and 3 V_{pp}.
The circuit may be powered by a 6 to 12 V supply: a PP3 battery (9 V) is perfect. Power consumption is about 48 mW.



067 FOUR-CHANNEL STEREO SWITCH

The circuit described here enables a choice to be made from four different stereo channels with only one switch. Internal switching is effected by CMOS devices to obviate crackling, bounce, and other annoyances associated with mechanical switches.
The two D-type bistables in IC_2 are connected as binary dividers by linking their $\overline{Q}$ output to the D input. The $\overline{Q}$ output of FF_1 is also linked to the clock input of FF_2, which results in a kind of four-bit counter.
The push-button is connected to the clock input of FF_1. The four OR gates, N_1 to N_4, decode the output states of the bistables, so that at all times only one gate has a high output.
The outputs of the gates drive the CMOS switches in IC_3 and IC_4. The outputs of the four electronic switches in these ICs are strapped together.
The input of each switch incorporates a potential divider, ensuring that the switches operate in their linear regions. This arrangement ensures minimum distortion of the audio signals: the negative parts of these signals would otherwise be distorted, since the switches work from an asymmetrical supply.
The circuit draws a current of only about 1 mA at a supply voltage of 5 V. The supply voltage may be increased to about 15 V.



"It's a pretty small battery-powered PROM programmer – so what?"

Tools which are convenient get used a lot – that justifies their existence. There is no way we could explain all the usefulness of S3 here. Instead, if you're interested we're going to let you see it, use it and evaluate it in your own workshop. We went to a lot of trouble to design S3 just the way it is – no other PROMMER is all CMOS and all SMT. So **we** must be convinced that S3 would be a formidable addition to your armoury. Now all we have to do is to convince **you**.

"Such a little thing can't be powerful, like a big bench-programmer – er – can it?"

Yes, it can. It is **more** powerful. S3 leaves other prommers streets behind. S3 has continuous memory, which means that you can pick it up and carry-on where you left off last week. S3 has a huge library EPROMS and EEPROMS. S3 can blow a hundred or more PROMS without recharging. **S3 also works remotely, via RS232.** There's a DB25 socket on the back. All commands are available from your computer (through a modem, even). Also S3 helps you develop and debug microsystems by memory-emulation.

"What's this memory-emulation, then?"

It's a technique for Microprocessor Prototype Development, more powerful than ROM emulation, especially useful for single-chip "piggy back" micros. You plug the lead with the 24/28 pin header in place of the ROM/RAM. You clip the Flying-Write-Lead to the microprocessor and you're in business. The code is entered using either the keyboard or the serial interface. Computer-assembled files are downloaded in standard format – ASCII, BINARY, INTELHEX, MOTOROLA, TEKHEX.
Your microprocessor can WRITE to S3 as well as READ. You can edit your variables and stack as well as your program, if you keep them all in S3.
S3 can look like any PROM up to 64K bytes, 25 or 27 series. Access is 100ns – that's really fast. Memory-emulation is cheap, it's universal and the prototype works "like the real thing".

S3 loads its working programs out of a PROM in its socket, like a computer loads from disk. Software expansion is unlimited. Upgrades will come in a PROM. Programs can be exchanged between users. How's that for upgradability?

"Can I change the way it works?"

You surely can. We keep no secrets. System Variables can be "fiddled." **New programming algorithms can be written from the keyboard.** Voltages are set in software by DACs. If you want to get in deeper, a Developers' Manual is in preparation which will give source-code, BIOS calls, circuit-diagrams, etc. We expect a lively trade in third-party software e.g. disassemblers, break-point-setters and single-steppers for various micros. We will support a User Group.

"I'm bound to let the battery go flat."

Quite so. But in practice it doesn't matter. S3 switches off after a half-hour of non-use anyway, or when the battery gets low. You don't lose your data. Then a slow-charge overnight or boost-charge for three hours will restore full capacity. You can keep using it when charging. So there really is no problem.

"I already have a programmer."

Pity it doesn't have S3 features, eh? But here's a trick worth knowing. If you plug S3's EMULead into the master socket of a ganger then you get an S3 with gang capacity. Isn't production separate from development anyway?

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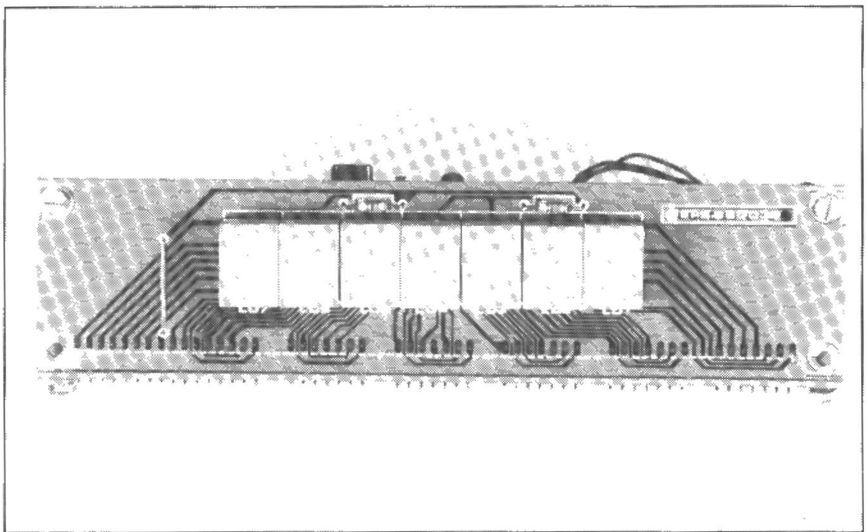
FREQUENCY READ-OUT FOR SW RECEIVER

Although the circuit described here was designed as a frequency read-out for a short-wave receiver, it can also be used as a universal counter.



The text of this article will be described on the basis of the 'SSB receiver for 20 and 80 metres' [1].

The block schematic of the unit is shown in Fig. 1. The signal to be measured is first fed to a variable-gain amplifier and then applied to an up/down counter via a NAND gate. The value at which counting starts depends on the setting of a number of preset switches. The final count depends on the setting of the up/down switch and may be larger or smaller than the preset value. To obtain a stable read-out, a latch has been introduced between the counter and the LED display, which stores the outcome of the measurement. The timing section serves to ensure that the other sections operate in time and with the required accuracy.

The block schematic of relevant sections of the receiver is given in Fig. 2 to show how the read-out unit transfers the information to the display. The signal whose frequency is to be measured is the output of the oscillator in the receiver, which is available at test point 2. To display the received frequency, the counter must start from 900,000, i.e., the IF of the receiver. Note that the counter counts in steps of 10 Hz. That value is set with the preset switches. Whether the counter is to count up or down depends on whether the received frequency lies above or under the IF. In the 20-m band, the IF is lower than the received frequency, and the counter must then count up from 900,000. In the 80-m band, the IF is higher than the received frequency, and the counter must then count down.

Technical Data

Bandwidth of input amplifier	1 kHz to 50 MHz
Input sensitivity	≧25 mV
Max. counter input frequency	25 MHz
Counter resolution	10 Hz
Gate time	0.1 s
Number of measurements per sec	5
Number of digits	7
Current drawn	≦465 mA

Preamplifier and timing section

The circuit of the input and timing sections is shown in Fig. 3. The preamplifier ensures that the oscillator in the receiver is not loaded unduly and raises the oscillator signal to a level of about 1 V_{pp} at the collector of T_4.

Preset P_1 serves to set the gain so that the counter operates in a stable manner, and also that T_4 does not go into saturation.

The amplified signal is fed to a Schmitt trigger, N_4, via C_3. Since the input of the trigger is at half the supply voltage level (R_3-R_4), the gate toggles readily at an input of 1 V_{pp}. Whether the output of N_4 is passed to the counter via clock N_3 is determined by the timing circuit, formed by IC_1, IC_2, and IC_3.

Circuit IC_1 contains an oscillator and a 14-bit divider. The oscillator frequency is determined by crystal X_1 (6.5536 MHz). The oscillator output is divided in IC_1 and IC_2 by 2^{17} to 50 Hz. This signal is divided by 10 in Johnson counter IC_3. The shape of the final 5 Hz signal is shown in Fig. 6, together with that of the signals at the other outputs of IC_3 and the outputs of N_1 and N_5.

The Johnson counter produces from each 50 Hz pulse five symmetrical square waves that have a PRF of 5 Hz and are separated from one another by 20 ms spaces.

After the first clock pulse, Q_1 goes high, and as long as it remains so (for 100 ms), the signal to be measured is passed from N_3 to the counter. Then N_1 transmits a pulse which ensures that the contents of the counter are passed to the latch, and thus to the display. To prepare the counter for the next measurement, N_5 transmits a pulse which is used to reset the counter, and this terminates the count cycle.

Table 1

Truth table Johnson counter IC3

Q1	Q2	Q3	Q4	Q5	state
0	0	0	0	0	0
0	0	0	0	1	1
0	0	0	1	1	2
0	0	1	1	1	3
0	1	1	1	1	4
1	1	1	1	1	5
1	1	1	1	0	6
1	1	1	0	0	7
1	1	0	0	0	8
1	0	0	0	0	9

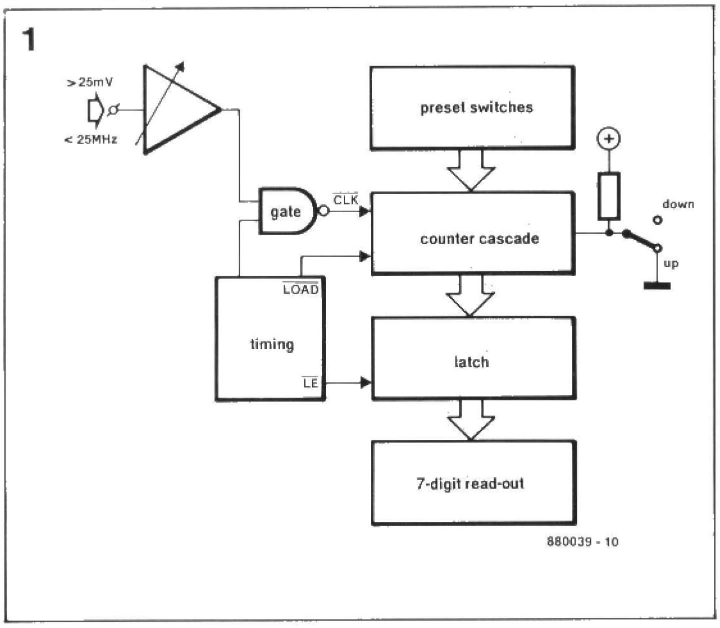


Fig. 1. Block schematic of the frequency read-out.

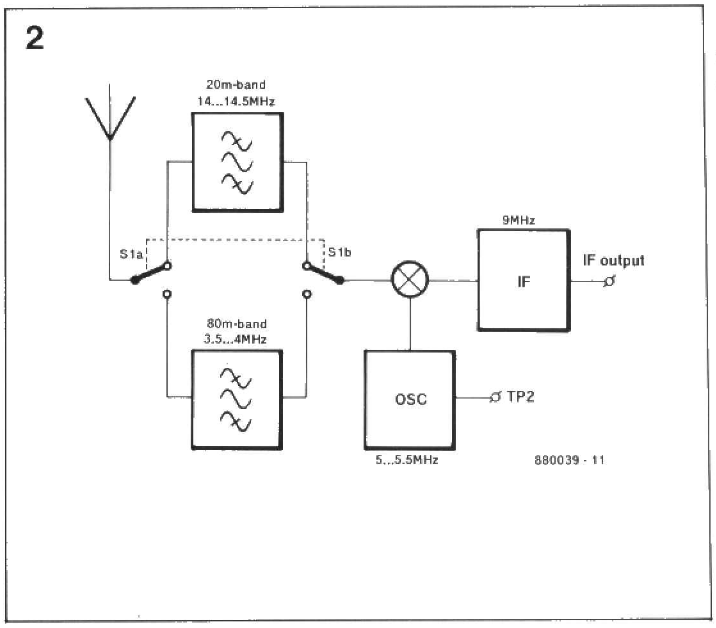


Fig. 2. Block schematic of the receiver parts relevant to the frequency read-out.

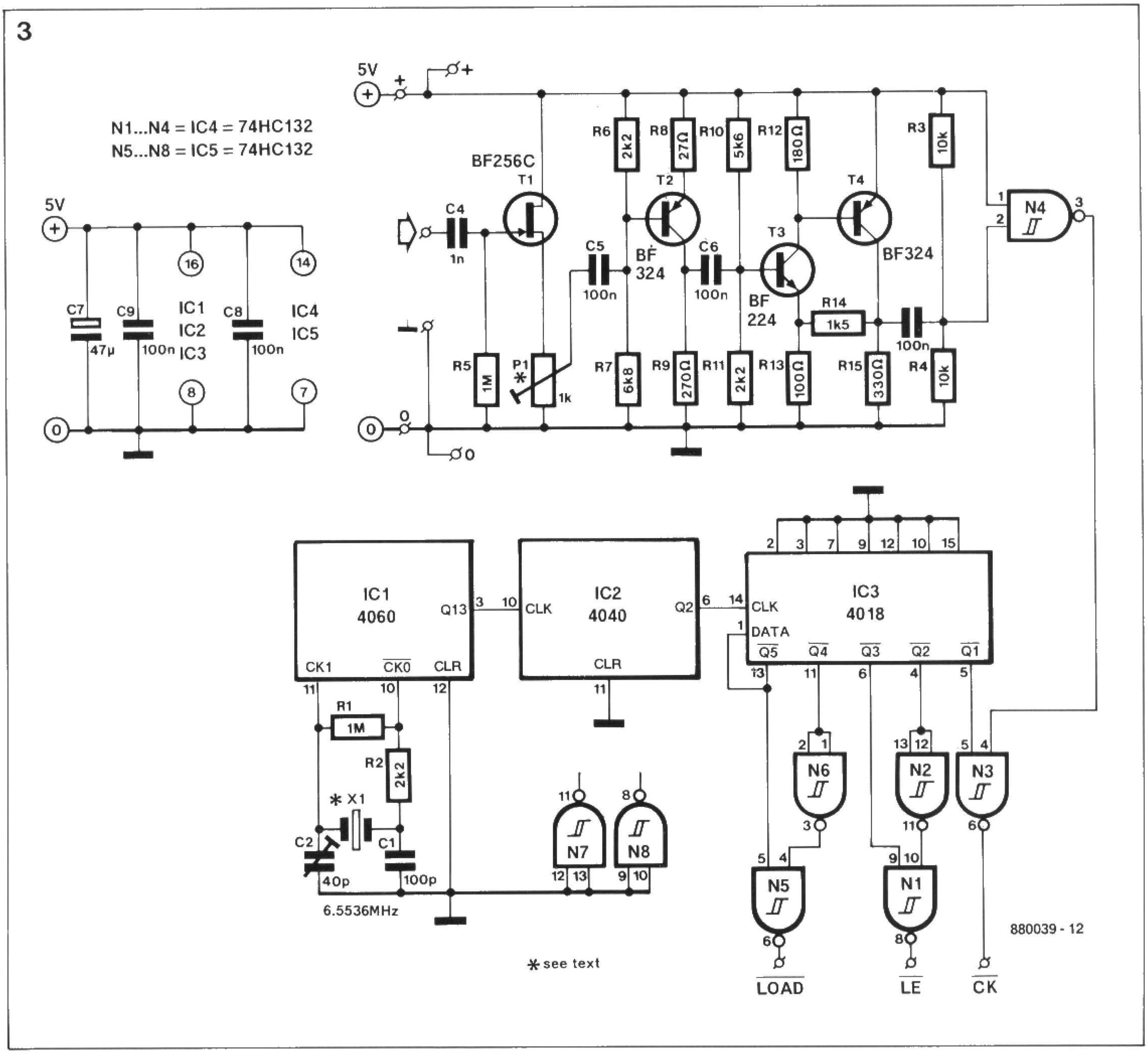


Fig. 3. Circuit diagram of the preamplifier and the timing section.

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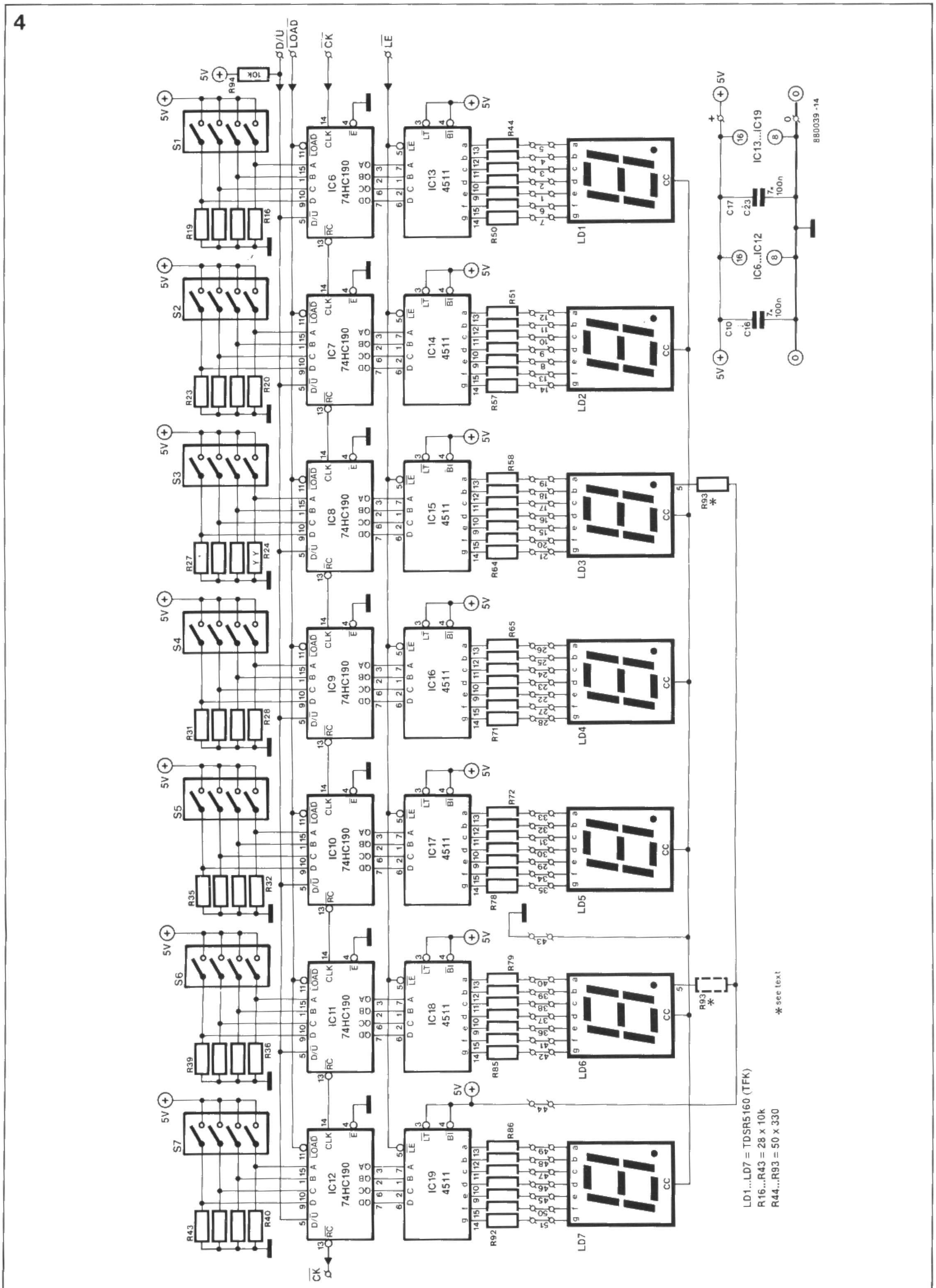

Fig. 4. Circuit diagram of the counter and display section, constructed from discrete components to reduce noise and other spurious emissions.

Parts list

Resistors (±5%):
R_1;R_5 = 1M0
R_2;R_6 = 2K2
R_3;R_4;R_{16}...R_{43} incl.;R_{94} = 10K
R_7 = 6K8
R_8 = 27R
R_9 = 270R
R_{10} = 5K6
R_{11} = 2K2
R_{12} = 180R
R_{13} = 100R
R_{14} = 1K5
R_{15};R_{44}...R_{93} incl. = 330R
P_1 = 1K0 preset H

Capacitors:
C_1 = 100p
C_2 = 40p trimmer
C_3;C_5;C_6;C_8...C_{23} incl. = 100n
C_4 = 1n0
C_7 = 47μ; 10 V; tantalum

Semiconductors:
T_1 = BF256C
T_2;T_4 = BF324 (Cricklewood)
T_3 = BF224 (Cricklewood)
LD_1...LD_7 incl. = D350PK or TDSR5160 (AEG-Telefunken; UK distributors are listed on InfoCard 502, *EE* February 1987). Suggested equivalents: HD1133R (Siemens*) or HDSP5303/5503/5603/5703/5733 (Hewlett-Packard).
IC_1 = 4060
IC_2 = 4040
IC_3 = 4018
IC_4;IC_5 = 74HC132
IC_6...IC_{12} incl. = 74HC190
IC_{13}...IC_{19} incl. = 4511

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Miscellaneous:
S_1...S_7 incl. = 4-way DIL switch block.
X_1 = quartz crystal 6.5536 MHz.
PCB 880039 (see Readers Services page).

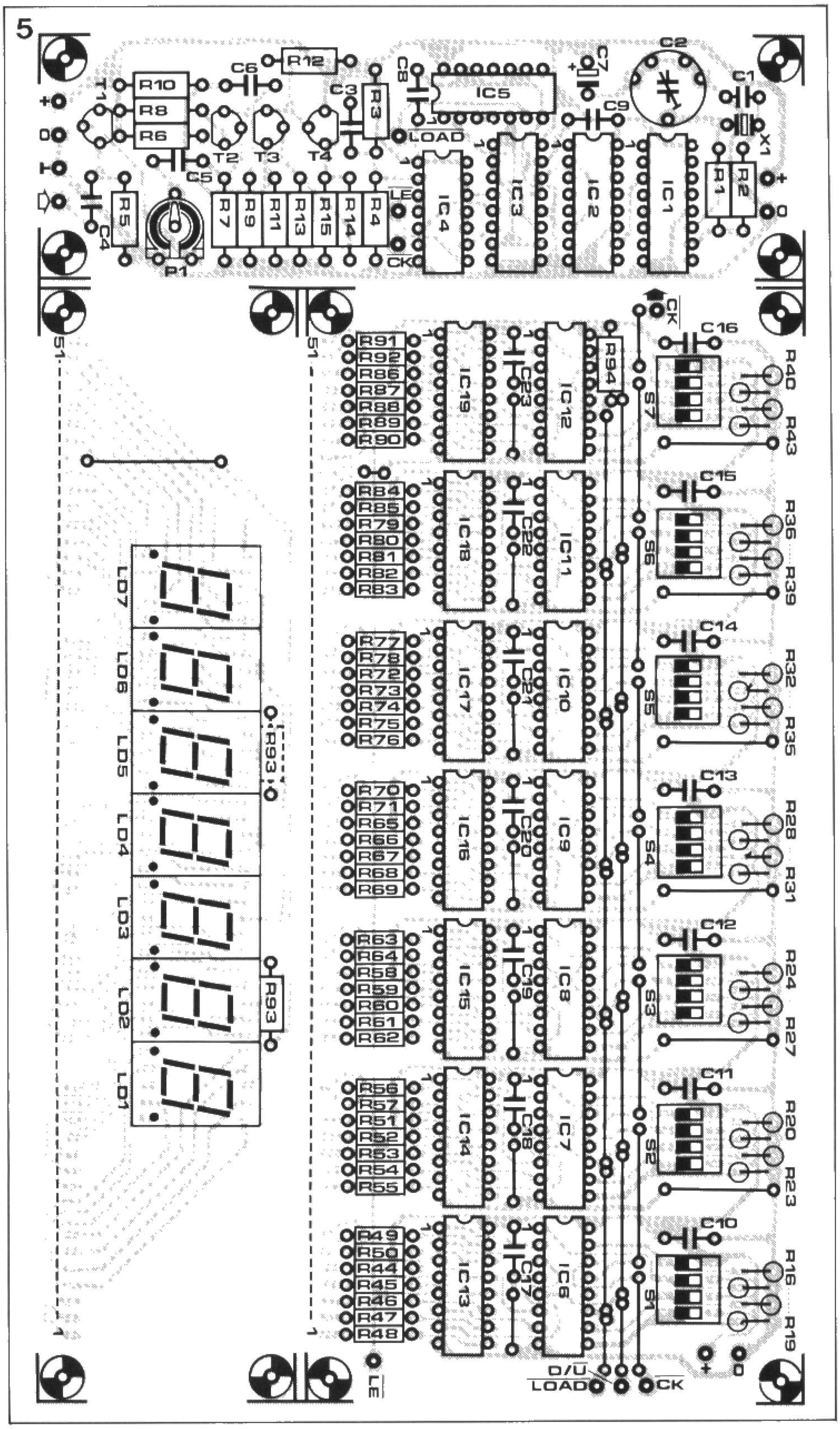


Fig. 5. The printed circuits of the three sections of the frequency read-out are contained on one board.

Discrete counting

The counter is constructed from discrete counter elements and display drivers (with built-in latch), as shown in Fig. 4. Each of the seven identical sections consists of four preset switches with pull-down resistors, a BCD counter, a BCD-to-seven-segment decoder/driver with latch, seven biasing resistors, and a seven-segment LED display.

Preset switches S_1 to S_7 set the BCD code to a value at which the counter starts counting. The pull-down resistors ensure that the preset inputs of the counter are provided with a 0 if the switch is open.

Counters IC_6 to IC_{12} incl. are coupled asynchronously. This has the advantage that only the first counter (IC_6) is supplied with a high clock frequency.

The result of the measurement is stored in latches IC_{13} to IC_{19} incl. These ICs ensure that the measuring result is made visible on the display, because they also have the BCD-to-seven-segment-decoder and a display driver on board. The position of the decimal point is determined by the location of R_{93}. If this is in the position shown solid, the display reads in kHz; in the position shown in dashed lines, the read-out is in MHz.

Construction

The three printed circuits are housed on one board, as shown in Fig. 5. The three sections should be cut off as required. The circuits are single-sided, which gives rise to a fair number of wire links. These

are, of course, best put in place first. Construction is facilitated by Fig. 7, which shows the layout of the prototype. Several types of switch may be used for S_1 to S_7. First, there are DIL switches, which are set according to the BCD code in Table 2. Note that the least significant bit is located at the left-hand side of the switch viewed from the display. Another, rather more luxurious, type is a BCD thumb-wheel switch: rather expensive, but very useful if the switches are set often. If the switches are used only once, they may be replaced by wire links. Resistors R_{16} to R_{43} and DIL switches S_1 to S_7 are then, of course, not required. Where a preset input must be 1, the wire link is fitted in place of the switch; if the preset input must be 0, the wire link takes the place of the relevant resistor. Take care that each preset input gets only one wire link — no more, no fewer.

The calibration of the read-out is carried out by turning P_1 so that its wiper is at ground potential, and then turning it until a stable read-out appears on the display. Take care that T_4 does not go into saturation with differing input signals at high frequency, because this will affect the bandwidth which may lead to erroneous measurements. Once P_1 has been adjusted correctly, the frequency of the crystal oscillator must be set with the aid of C_2. This is done most conveniently by tuning the receiver to a strong station whose frequency is known accurately: C_2 is then rotated until that frequency is displayed.

Table 2

	BCD code			
	D	C	B	A
0	0	0	0	0
1	0	0	0	1
2	0	0	1	0
3	0	0	1	1
4	0	1	0	0
5	0	1	0	1
6	0	1	1	0
7	0	1	1	1
8	1	0	0	0
9	1	0	0	1

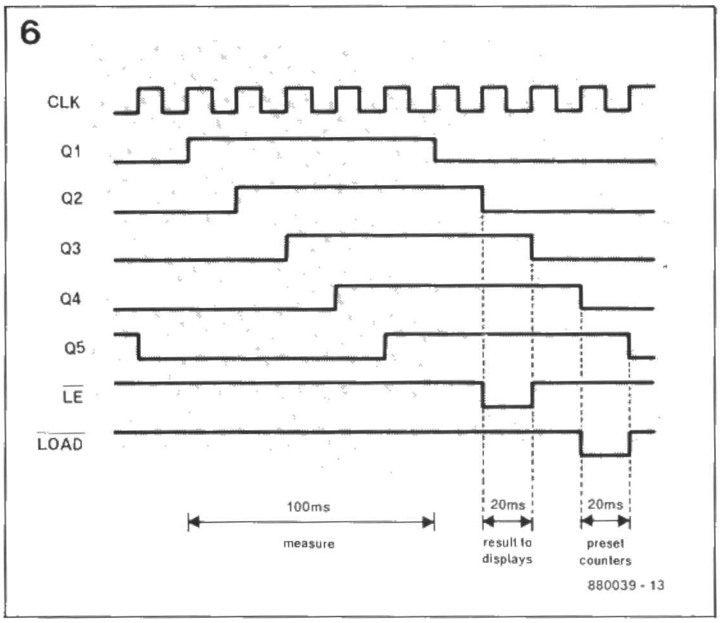


Fig. 6. Output signals of IC_3, N_1, and N_5.

Accuracy

The accuracy of the read-out unit is determined mainly by crystal X_1. Without special precautions, the frequency of the received signal is measured with an accuracy of within 200 to 300 Hz. The short-term accuracy may be improved by placing the crystal in polystyrene foam and tuning the oscillator (before any measurement) with the aid of a reference oscillator. This will guarantee good accuracy for up to an hour.

If greater accuracy and stability are required, a crystal with a temperature coefficient of 20 to 30 ppm should be used. Bear in mind that the effect of temperature becomes greater at high frequencies.

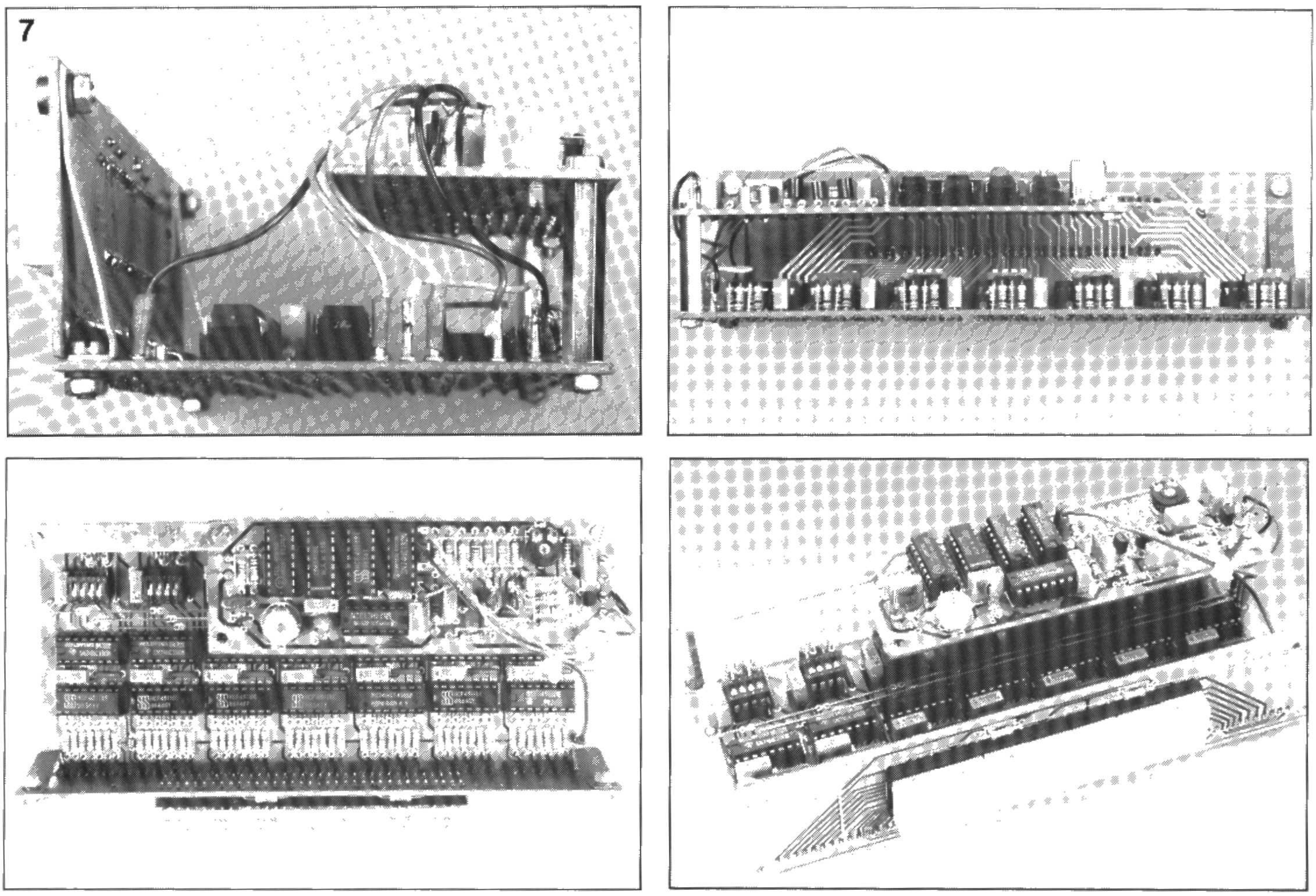


Fig. 7. Various views of the prototype frequency read-out.

Finally

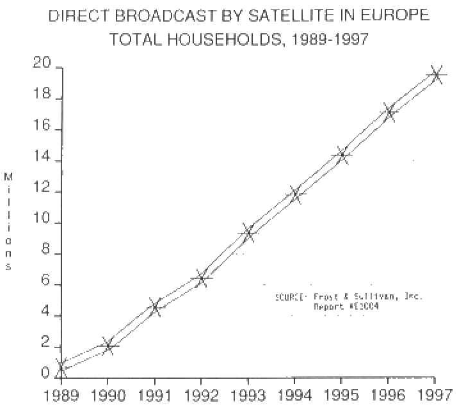
The read-out lends itself readily to experimenting. It is, for instance, possible to use an external and very stable frequency source, such as the 'Intelligent Time Standard' [2] to enable measurements to be made with a resolution of 1 Hz. The (divided) external oscillator signal is connected across pins 6 and 8 (ground) of the socket for IC_2 (IC_1 and IC_2 and associated components are, of course, nor required and may be omitted or removed). The frequency, f_i, of the (divided) external signal is determined by $f_i = 5/T_m$, where T_m is the measuring time (=1 s for a frequency of 1 Hz).
With these changes in the timing section, it may, of course, happen that it is no longer possible to place the decimal point at its correct position with R_{93}. In that case, the resistor will have to be mounted on the display PCB (track side) between the +5 V track and pin 5 of the relevant display module.
Speed is another aspect with which may be experimented. The maximum frequency the counter can handle is determined by the highest clock frequency to which IC_6 can react. The first thing to do is to try each of the seven modules in the IC_6 position: in the prototype one or two of the modules worked satisfactorily up to 25 MHz, although the data sheet indicated 17 MHz (Type 74190). The modules shown in Fig. 4 (Types 74HC190) can operate with clocks of up to 40 MHz. If even higher speeds are required, use a Type 74F190 in the IC_6 position, which is intended for operation with clock frequencies of around 90 MHz. The other ICs can remain HC types, because they need not work at these high frequencies. It is also possible to use HCT types: these are essential, by the way, if it is required to drive the counter from a TTL circuit.
It should be noted that the design of the counter circuit with discrete components is deliberate: a multiplexed integrated display unit is a source of noise and interference, which in a sensitive receiver is, of course, the last thing you want. None the less, it is still advisable to mount the read-out in a properly screened enclosure.

References:
(1) *Elektor Electronics,* November 1987, pp. 52-56.
(2) *Elektor Electronics,* February 1988, pp. 22-30.

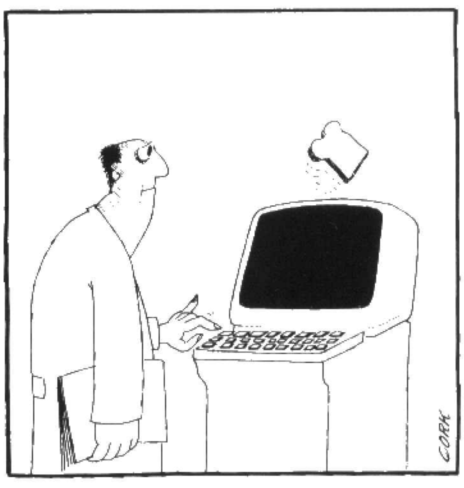
RADIO & TV NEWS

Direct Broadcast by Satellite

Direct Broadcast by Satellite—DBS—in western Europe is gathering momentum. Its enormous commercial potential, the technology involved, advantages and disadvantages against competing systems, the regulatory environment, risks and insurance are examined in a new study from Frost & Sullivan: ***Direct Broadcast by Satellite in western Europe*** (#E1004).
The first DBS satellite, TV-SAT 1, was developed by the Eurosatelllite consortium and launched in November, 1987. Eurosatellite is a French-German owned company. The next launch planned is France's TDF-1, followed by Luxembourg's medium-powered ASTRA satellite, Sweden's DBS satellite Tele-X and others, including a DBS satellite by the British Satellite Broadcasting (BSB) consortium.
For 1989, the first full year of European DBS and medium -powered satellite television, the Frost & Sullivan study forecasts a total audience of 700,000 households—rising to almost 19.5 million by 1997. The yearly growth rate, over 100% initially, will taper off to 28% on average between 1991 and 1997. Of course, audience growth will not be uniform across Europe. The already high level of cable penetration in Belgium and the Netherlands, for example, makes them barren territory for DBS. The U.K. is expected to be the largest DBS market in terms of both subscription revenue and home-receiver product revenue.
Frost & Sullivan sees subscription revenue from DBS and medium-powered satellite television in Western Europe soaring from $7.9 million in 1989 to over $702 million by 1997 (fixed 1987 U.S. dollars). The U.K., with annual



subscription revenue of $190 million by 1997, will account for more than a quarter of the total. France ($147.6 million) and West Germany ($150 million) will each account for about a fifth. In addition, advertising revenue in Western Europe from satellite television is forecast to go from $7.3 million in 1989 to $608 million by 1997.
Looking at product revenue from the sale of receivers and outdoor units, the study predicts a rapid rise from $111 million in 1989 to a peak of $684 million by 1992, of which $424 million will come from receiver sales and $260 million from sales of outdoor units. The French product market, however, will be slower to take off and will peak later at $229 million in 1996. Product revenue will peak in 1992 at $362.4 million in the U.K. and at $116.9 million in West Germany. Annual product revenue overall in Western Europe will fluctuate from 1992 to 1997 above $400 million.
The study points out that DBS for all its potential, is "merely a delivery technology" competing with other delivery systems: terrestrial television and cable systems, as well as low and medium-powered satellites. Cost factors certainly play a large part in this increasingly competitive market, but Frost & Sullivan predicts that DBS's success will depend ultimately on the range and quality of its programming.
This report compares in detail DBS and other systems, covering all aspects from political/regulatory constraints to the dramatic reductions in size and cost of dishes for receiving satellite television. Good reception can be achieved with dishes as small as 35 centimeters in diameter, costing around $320. Included in the study are analyses of equipment suppliers and industry structure.
Each of the DBS projects, as well as other satellite projects, is examined in depth as are the markets in various countries. The U.K. will likely be the most competitive satellite TV market. Luxembourg's ASTRA, due to go up in 1988, will have eight English channels. The U.K.'s own project (privately financed and operated), the BSB satellite, is scheduled for launch in 1989, providing three channels. Then, in 1990, Atlantic Satellites Ltd. plans to lauch an Irish satellite, with five channels, competing for the U.K. market.



GaAs FET CONVERTER FOR 23 CM AMATEUR TELEVISION

The ultra-low noise figure of this SHF converter, in conjunction with its high conversion gain, enables it to outperform almost any combination of a 23 cm preamplifier and ATV converter based on bipolar transistors and a passive mixer respectively.

from an idea by G. Wehrhahn DD9DU

Transmission and reception of fast-scan television is among the most exciting and technically interesting modes available to the licenced radio amateur with an interest in UHF and SHF design and construction. The 70 cm band (430—440 MHz) has long been popular for amplitude-modulated ATV with vestigial sideband suppression, but its use for ATV (a wideband mode) is now under considerable pressure owing to the rising demand for frequency allocations for narrow-band and professional services (radar, mobile telephones, cellular radio). Forced to 'move up', ATVers started to explore the 23 cm band, in which the frequency range allocated to ATV (1250—1285 MHz) was found to offer plenty of space for AM as well as a new mode, wideband FM ATV. This mode, initially tried and used by French ATVers, allows transmitter output stages to operate in class C rather than A or AB, freeing the station operator from the output power restrictions imposed by the required ultra-linear operation of the entire transmitter, from video input to RF output. Hence, FM facilitates design of TV transmitters in general and semiconductor-based power output stages with acceptable efficiency in particular. The only disadvantage with respect to AM is the higher bandwidth (15—25 MHz instead of 7 MHz), but this is not generally a problem in the 23 cm band where narrow-band telephony is allocated 'up-band' between 1296 and 1298 MHz. A further interesting aspect of 23 cm ATV is that commercial equipment for this mode is hardly available.

Television relay stations

Over the past few years, the use of the 23 cm band for FM ATV has gained considerable popularity thanks to many groups of dyed-in-the wool ATVers all over the UK and Europe building sophisticated relay stations (*repeaters*) installed in elevated locations. Repeater power output is usually between 5 and 10 watts, and the normally usable range between 15 and 50 km. There exist cross-band repeaters (70 cm AM input — 23 cm output, AM or FM) as well as in-band repeaters (1275 MHz FM input — 1252/1310 MHz AM output, or the other way round).

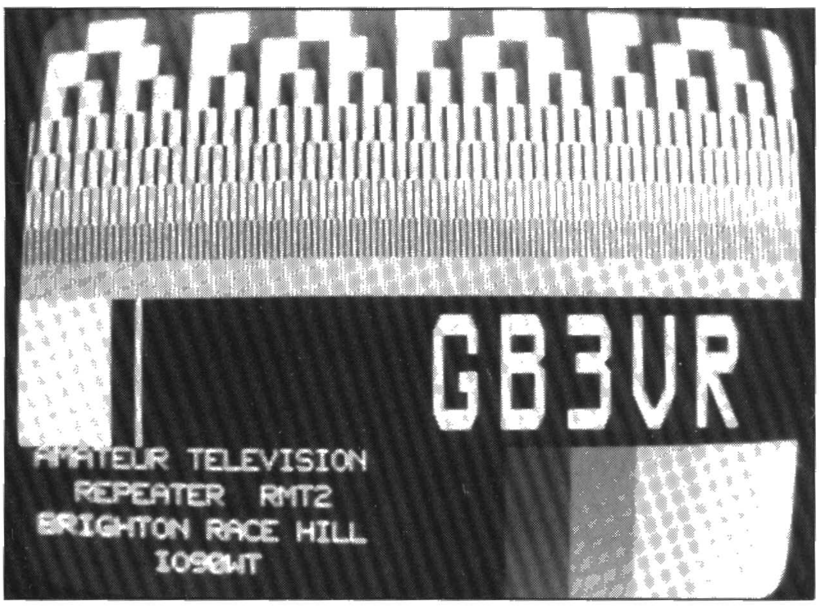
During periods of favourable propagation (*lift conditions*), reception of 23 cm ATV repeaters — and, of course, privately owned and operated ATV stations — has been reported over distances up to 150 km.

ATV repeaters are so important because most of them have omnidirectional aerials, and can be remote-controlled to operate as a beacon transmitting a test chart. This facility enables a strong, steady 23 cm ATV signal to be used for setting up, comparing and testing aerials and converters locally.

In most countries, ATV signals may be received without a licence or permit. It may, however, be necessary to obtain a permit for installing the aerial. Because of the line-of-sight propagation of signals in the 23 cm band (1250—1300 MHz), the aerial should be mounted as high as possible without running into excessive cable losses.

The construction and operation of a 23 cm transmitter or receiver represents a radical departure from established techniques on the shortwave bands, and for this reason the construction of the GaAs FET converter is recommended only to those who have experience working with microwave circuits.

Many of our readers are avowed radio amateurs, or have a general interest in the radio hobby. This fact obviates the need for an introduction into SHF techniques in general and fast-scan amateur



Test chart of GB3VR, the Worthing & District ATV repeater located at Race Hill, Brighton, QTH locator IO90WT.

television on 23 cm in particular, which would be outside the scope of this article in any case. Readers unfamiliar with these subjects will find useful information and references at the end of this article. ATV stations can be found around 1252 MHz (AM) and 1275 MHz (FM and repeaters).

A number of national and international contests are organized every year to boost the interest in ATV, and, of course, to enable participants to compete for the highest score measured as the total number of kilometres covered by the station. ATV activity is highest during these contests, and newcomers are recommended to listen on to the international ATV calling channels, 144.750 MHz (mode: FM) and 144.170 (mode: USB) in the 2 metres band. Almost any operator of an ATV transmitter welcomes a reception report, and the fact that a contest is going on does not, in general, mean that there is no time and patience for sending a picture to a receive-only station ready to align a newly built converter. Contrary to those oganized for telephony stations, ATV contests are usually held in an atmosphere of light-hearted competition with very few stations not keeping to the rules.

The converter described here is the perfect introduction to 23 cm ATV because it is a relatively inexpensive and simple design. It has only one preamplifier stage, an active mixer, and a free-running, single-transistor local oscillator. Construction is also fairly straightforward thanks to the use of a small PC board with printed inductors (micro-striplines).

All prototypes of the GaAs FET converter were found to give better results than a formerly used combination of a tuned two-stage stripline preamplifier for 23 cm using (very expensive) bipolar transistors Type NE64535 from NEC, and a 23 cm converter based on a crystal-controlled local oscillator chain and a Schottky diode mixer (designed by DJ5XA and described in edition 2/1975 of *VHF Communications*).

Interestingly, the cost of the GaAs FET converter is much lower than this (now technically outdated) combination of a preamplifier and a converter.

Only three transistors

The coming of gallium-arsenide semiconductors has enabled receiver noise figures to fall below values that are virtually impossible to achieve with bipolar transistors available to the radio amateur. The GaAs FETs used in the present converter are relatively inexpensive dual-gate types Type 3SK97. Types S3030 (Texas Instruments) and CF300 (Siemens) were also tried with excellent results. Contrary to the popular belief, there is nothing mysterious about GaAs FETs: in fact, their outlook, static and dynamic operation is very similar to that of well-known VHF or UHF dual-gate MOSFETs in the 3Nxxx and BF9xx series. The main advantage of the GaAs FET used here is that it can offer an in-circuit noise figure that remains below 2 dB for frequencies up to 1.5 GHz. Furthermore, gain is high but stable, and matching to tuned circuits is fairly simple thanks to the extremely low internal capacitance that results in a small reactive component.

The circuit diagram of Fig. 1 shows that the aerial signal reaches gate 1 of pre-

1

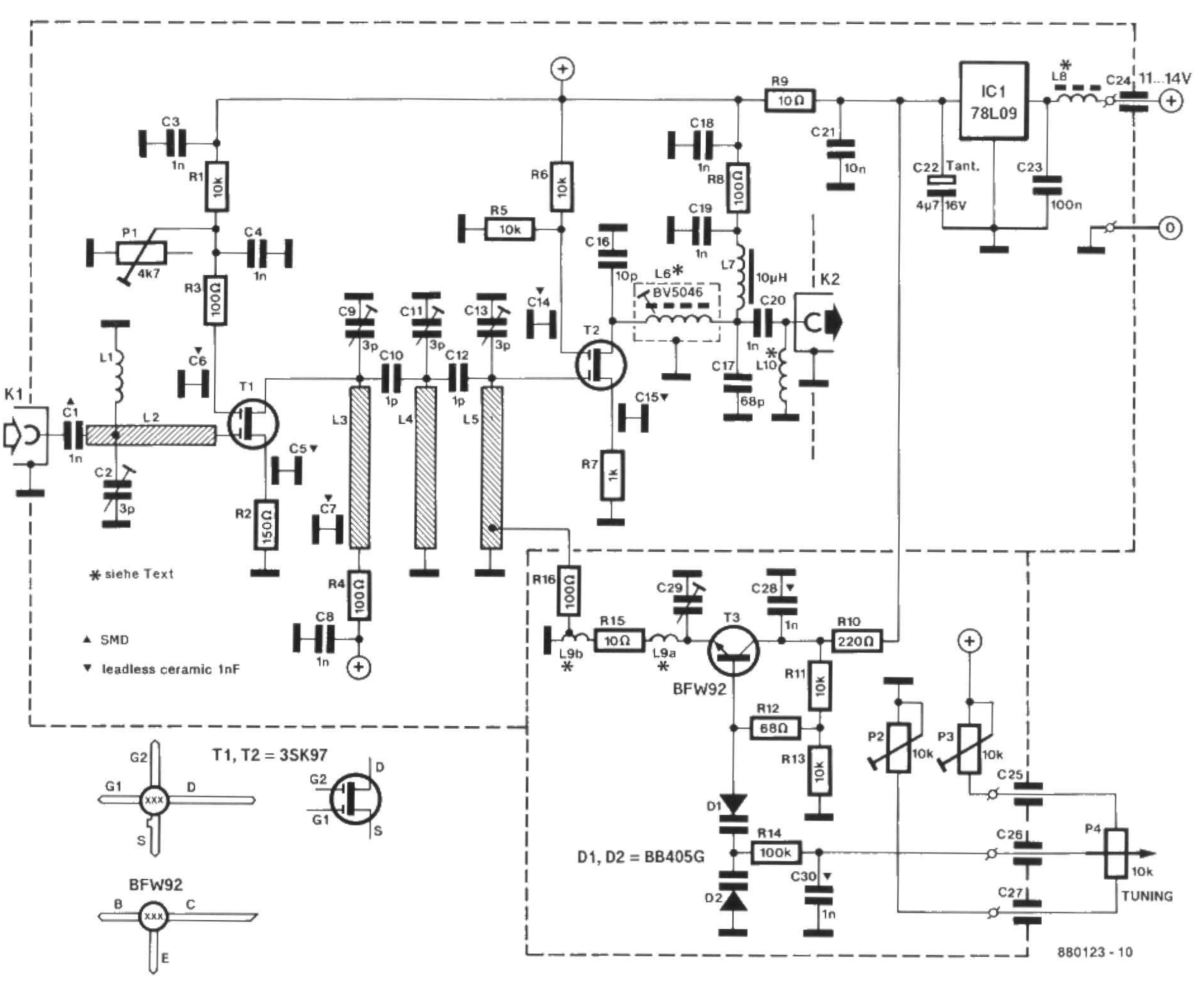


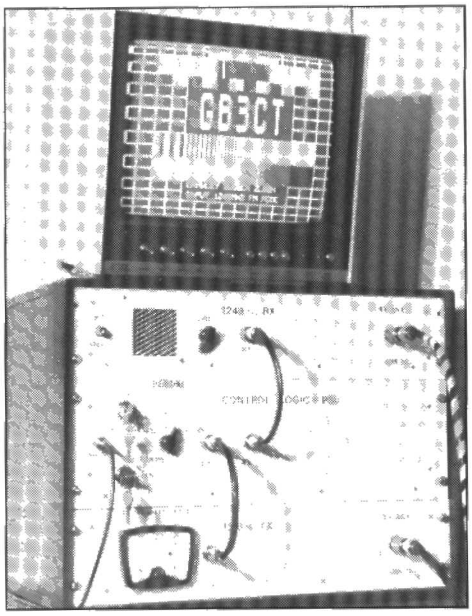
Fig. 1. Circuit diagram of the converter for amateur television reception in the 23 cm band.

amplifier T_1 via micro-stripline inductor L_2. Matching of the transistor to the cable impedance of 50 Ω is optimized by adjusting trimmer C_2. Preset P_1 allows adjusting the drain current of T_1 to obtain either high gain or a low noise figure. In most cases, a compromise between these will have to be found.

The amplified 23 cm signal is passed to mixer T_2 via a three-element top-coupled micro-stripline filter, which is tuned by means of trimmers. It should be noted that C_{10} and C_{12} increase the total bandwidth of the filter to a value suitable for reception of 27 MHz wide FM ATV. For AM ATV, these capacitors may be omitted to achieve pure inductive coupling resulting in lower bandwidth.

The local oscillator signal reaches gate 1 of T_2 via R_{16} and a low-impedance tap on L_5. The intermediate output frequency of the converter can be chosen freely between 40 and about 200 MHz. In prototypes, the drain circuit of T_2 was tuned to 48 MHz by C_{16}-L_6-C_{17} to enable the converter to be used for AM ATV reception with a portable colour TV set tuned to VHF Channel 2 (now no longer used for TV in the UK). Provided C_{16}-L_6-C_{17} and L_{10} are dimensioned accordingly, the intermediate frequency is simple to move up to, say, 180 MHz (Channel 6 in Band 3). Obviously, the higher the intermediate frequency, the better the image rejection of the mixer.

A domestic television set is, of course, not suitable for receiving frequency-modulated ATV, which is gaining popularity because of the benefits already mentioned. For FM ATV, a special intermediate frequency amplifier will have to be made, followed by a wideband FM demodulator. The most commonly used IF for FM TV is 70 MHz, but here, again, the IF frequency can be chosen freely.



A fine piece of advanced electronics built by radio amateurs: the Crawley ATV repeater, GB3CT.

The single-transistor, varicap-tuned, local oscillator is a slightly modified version of that discussed in Ref. 1. Properly constructed, its stability is so good that an AFC circuit is not required. Presets P_2 and P_3 enable defining the tuning range of the converter. Capacitor C_{29} is a coarse frequency adjustment, and also serves to stabilize the power output of the oscillator. This trimmer, which may not be needed in all cases, is simply 10 mm or so of straight wire positioned above the PCB surface (read the part on C_x in Ref. 1). Although not apparent from the circuit diagram, the actual length of the anode lead of D_1 and the construction of L_{9a} also determine the frequency of oscillation. The oscillator can be set to operate roughly between 1000 MHz and 1500 MHz.

Finally, the dashed lines in the circuit diagram denote a screen around the local oscillator to prevent stray radiation.

Construction

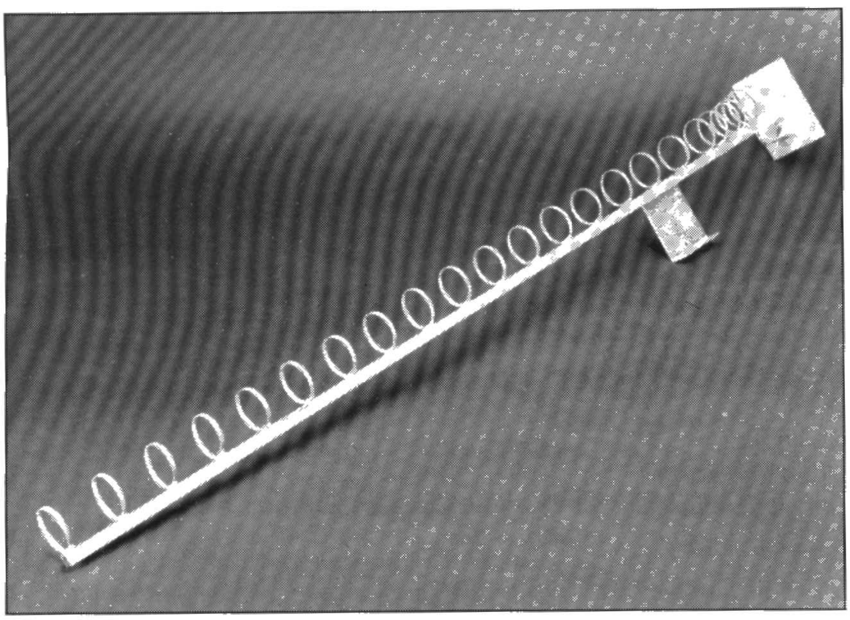
Figure 2 shows the printed circuit board designed for the converter. In de description below, the upper drawing is called the *component side* of the board, and the lower drawing the *reverse side* (*soldering side* would be incorrect because a number of components are also soldered at the component side). Construction is fairly simple for those grown accustomed to the use of the leadless ceramic capacitors. The actual value of these is uncritical (anything between 470 pF and 1.5 nF will work; 1 nF is the most commonly available value). There are 7 of these capacitors in the converter — each is fitted vertically in a slot which is carefully jig-sawed or drilled and filed in the PCB. The length of the slots is such that the shoulders of the leadless capacitors rest on the PCB surface. The holes for the two GaAs FETs are drilled to 3.5 mm. T_3 is not fitted in a hole.

The cross-sectional views of the PCB in Fig. 3 show the connections of the gate 2 and source terminals of T_1 to decoupling capacitors C_6 and C_5 respectively. Micro-stripline L_4 is connected to ground by a small piece of copper foil. All 1 nF capacitors (and C_{21}) not marked with a black triangle in the circuit diagram are miniature ceramic types with a lead spacing of 2.5 mm.

Input inductor L_1 is one turn of 0.5 mm dia. silver-plated wire. Choke L_8 is wound as 6 turns of 0.2 mm dia enamelled copper wire through a ferrite bead or small balun.

Inductors L_{9a} and L_{9b} are formed by the wire terminals of R_{15}. L_{9a} is 2 turns with an inside diameter of about 3 mm and a turns spacing of 1 mm. The other inductor, L_{9b}, is the straight-wire terminal soldered to ground, as shown on the component overlay. A 2 mm hole is made in the screen surrounding the local oscillator, so that R_{16} can be soldered to a tap on L_{9b}, approximately 10 mm from where this is bent down and connected to ground. It is important that R_{15} runs horizontally at about 4 mm above the PCB surface. Also make sure that it does not cause excessive strain on the emitter lead of T_3.

When required, coupling capacitors C_{10} and C_{12} are fitted direct onto the micro-



The most commonly used directional aerial for 23 cm is the so-called *loop-yagi*. This photograph shows a 24-element type scaled for 2304 MHz (13 cm).

2 **Note: only T1, T2 and C1 are mounted at this side of the board**

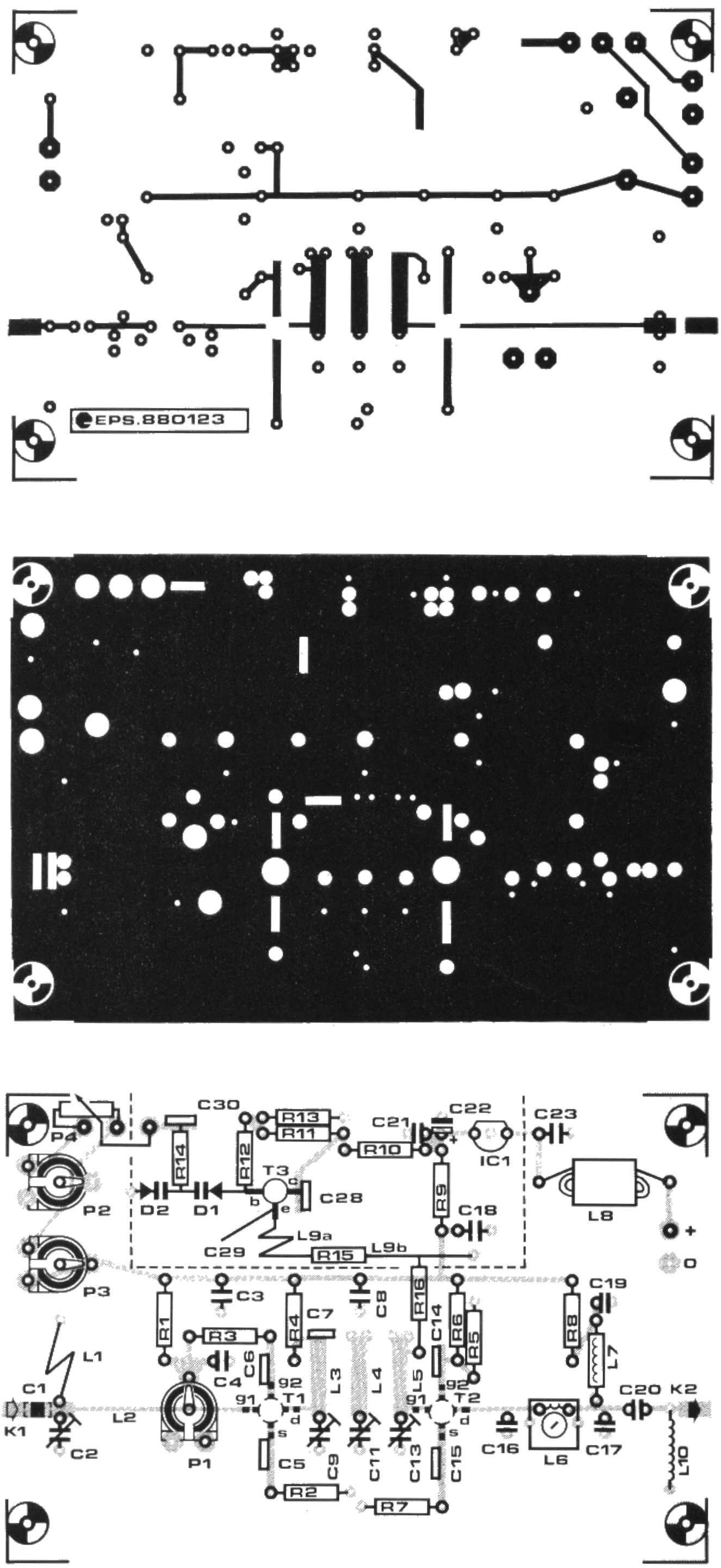


Fig. 2. Double-sided printed circuit board for the converter.

Parts list

Resistors (0.25 W carbon film; ±5%):
R1;R5;R6;R11;R13 = 10K
R2 = 150R
R3;R4;R8;R16 = 100R
R7 = 1K0
R9;R15 = 10R
R10 = 220R
R12 = 68R
R14 = 100K
P1 = 4K7 or 5K0 preset H
P2;P3 = 10K preset H
P4 = 10K linear potentiometer

Capacitors:
C1 = 1n0 chip or SMD (Bonex; VeroSpeed; Cirkit)
C2;C9;C11;C13 = 3p subminiature trimmer (manufacturer: Sky) (C-I Electronics)
C3;C4;C8;C18;C19;C20 = 1n0 ceramic
C5;C6;C7;C14;C15;C28;C30 = 1n0 leadless ceramic (Cirkit; Bonex)
C10;C12 = 1p0 (see text)
C16 = 10p
C17 = 68p
C21 = 10n ceramic
C22 = 4μ7; 16 V; tantalum
C23 = 100n
C24 . . . C27 incl. = 1n0 feedthrough (solder type) (Cirkit; Bonex)
C29 = see text.

Inductors:
L1 = see text.
L2;L3;L4;L5 = micro-stripline on printed circuit board.
L6 = Neosid BV5046 (yellow-blue; 0.9 μH; 5 . . . 50 MHz) (C-I Electronics).
L7 = 10μH axial choke.
L8 = see text.
L9 = see text.
L10 = see text.

Semiconductors:
D1;D2 = BB405G (Bonex; C-I Electronics)
IC1 = 78L09
T1;T2 = 3SK97 (C-I Electronics)
T3 = BFW92 (Cirkit)

Miscellaneous:
K1;K2 = BNC socket (flange type).
PCB Type 880123 (not available ready-made through the Readers Services).
Tin-plate box with top and bottom lids. Size: 111 × 74 × 50 mm.

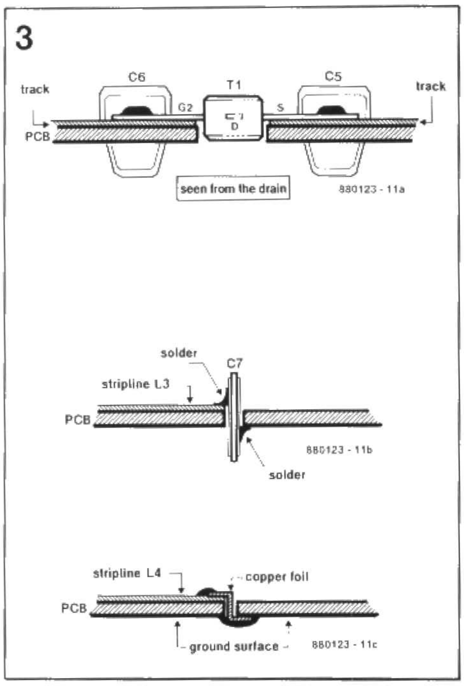


Fig. 3. Showing the use of leadless ceramic capacitors on the PCB (3a; 3b), and the connection to ground of micro-stripline L_4 with the aid of a small piece of copper foil (3c).

striplines, keeping the leads shorter than 1 mm.

The GaAs FETs are the last parts to be mounted on the PCB. As they are very small and static-sensitive, soldering must be done fast, with utmost care and using a low-power iron with a grounded tip.

The completed PCB is mounted in a tin-plate box with feed-through capacitors for the direct voltages and holes for BNC sockets K_1 and K_2. These are positioned such that the centre pin can be soldered direct on to the copper area provided. The PCB edges at the reverse side of the board are soldered direct to the inside of the box panels. When a ready-made tin-plate box is not available, an alternative enclosure can be made from cut-to-size pieces of unetched printed circuit board.

Setting up

The simplest way of aligning the converter is to ask for the assistance of a radio amateur licenced to transmit ATV on 23 cm. Alternatively, in the UK, get in contact with the BATC (British Amateur Television Club) to find out if there is a repeater within 20 km or so of your home. For the following description, it is assumed that a 23 cm AM ATV signal is available, and that the converter is used in conjunction with a TV set tuned to Channel 2.

Set all presets and trimmers to the centre of their travel. Adjust P_1 for a drop of 1.3 V across R_2. Check that oscillator works by measuring the drop across R_7; short-circuit the emitter of T_3 to ground to stop oscillation. This should cause the drop across R_7 to fall by about 0.2 V. Peak L_6 for maximum noise output of the converter. Then tune to the ATV signal, and peak the trimmers for optimum reception. This is fairly easy when the signal is relatively strong initially. Reduce the signal strength by carefully turning away the receive aerial, and redo all adjustments for best reception. It may be necessary to bend C_{29} closer to the PCB, or space the turns of L_{9a} wider, to stabilize LO output across the selected tuning range. Note, however, that re-positioning C_{29} changes the LO frequency, so that the tuning control, P_4, must be operated to restore reception. Also, P_2 and P_3 may have to be readjusted to obtain the correct tuning range.

Reference:

1. Indoor Unit for satellite TV reception, Part 1. *Elektor Electronics* October 1986.

For further reading:

Design and operation of the loop-yagi aerial: *VHF/UHF Manual*, by D.S. Evans G3RPE, and G.R. Jessop G6JP. Published by the Radio Society of Great Britain (RSGB).

Information on fast and slow-scan amateur television is available from the British Amateur Television Club • Mr. Dave Lawton G0ANO • "Grenehurst" • Pinewood Road • High Wycombe • Bucks P12 4DD.

23 cm ATV repeaters in the UK

callsign	location	channel	contact
GB3UT	Bath	RMT-1	G4JQP
GB3VI	Hastings	RMT-1	G3ZFE
GB3CT	Crawley	RMT-2	G4ZPP
GB3GT	Glasgow	RMT-2	GM1FGO
GB3RT	Rugby	RMT-2	G6IOM
GB3GV	Leicester	RMT-2	G4EUF
GB3PT	Cambridge	RMT-2	G8XMS
GB3TV	Dunstable	RMT-2	G4ENB
GB3UD	Stoke-on-Trent	RMT-2	G4DVN
GB3VR	Brighton	RMT-2	G4WTV
GB3AF	Newcastle-upon-Tyne	RMT-2	G1FBY
GB3HV	High Wycombe	RMT-3	G4CRJ
GB3ZZ	Bristol	RMT-2	G8VPG
GB3ET*	Emley Moor	RMT-2	G8CJS
GB3NV*	Nottingham	RMT-2	G8BWC
GB3HL*	Hull	RMT-2	G4HJD

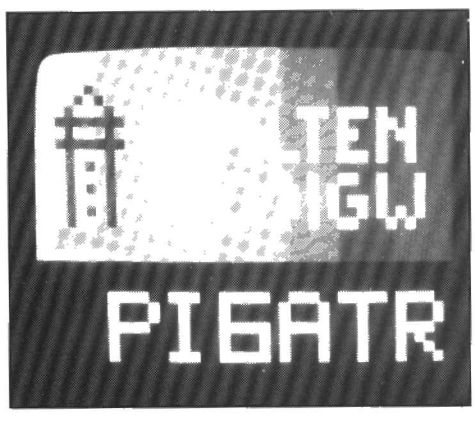
* not yet operational or awaiting licencing.

RMT-1 = input 1276.50 MHz (FM/AM)	output 1311.50 MHz (AM)
RMT-2 = input 1249.00 MHz (FM)	output 1318.50 MHz (FM)*
RMT-3 = input 1248.00 MHz (FM)	output 1308.00 MHz (FM)
ATV simplex (Europe): 1255 MHz	

* under review and likely to be moved to a lower frequency.

Information reproduced by courtesy of *CQ TV Magazine*.

Readers outside the UK should contact their national Radio Amateur Society or Societies for further information on ATV.



Black-and white reproduction of the colour test chart transmitted by PI6ATR, the first operational in-band FM/AM compatible 23 cm ATV repeater in Aalten, the Netherlands (QTH locator JO31GW). The transmitter supplies about 10 W in an omnidirectional bat-wing aerial fitted on the balcony of the 50 m high local water-tower. PI6ATR was built by PA3AOG, PA2AAD, PE1CHY, NL5184 and PA2ENG. The station was recently upgraded with a 70 cm AM ATV input.

I/O EXTENSION CARD FOR IBM PCs AND COMPATIBLES

(Part 2)

The concluding instalment deals mainly with the construction and testing of the I/O extension card, and also gives useful hints for programming.

Table 4 shows how the logic level at the control input of the DAC selects between two ways of dividing the twelve bits between two bytes (CTRl = Ø: left justified data; CTRL = 1: right justified data). Digital-to-analogue conversion is started when new data is written to the least significant byte.

Table 4.

	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0	D_7	D_6	D_5	D_4	D_3	D_2	D_1	D_0
ctrl = 0 →	B_{11}	B_{10}	B_9	B_8	B_7	B_6	B_5	B_4	B_3	B_2	B_1	B_0	x	x	x	x
ctrl = 1 →	x	x	x	x	B_{11}	B_{10}	B_9	B_8	B_7	B_6	B_5	B_4	B_3	B_2	B_1	B_0

Circuit description

The circuit diagram of Fig. 7 is really quite straightforward because of its resemblance to the previously discussed block diagram (see Part 1). Bus buffer IC_2 is an inverting type to save on ICs. Inverted address lines are applied to PAL (Programmable Array Logic) IC_4, which is programmed to provide all chip select signals required on the card.

Ports A and C in PPIs IC_5 and IC_6 are wired to the I/O connector. Port B is used for generating control signals for a number of chips on the card. Binary counter IC_7 derives the clock signal for the PIT from the system clock. DIL switches $S_1 \ldots S_5$ connect one of four counter outputs, or the the system clock, to the clock inputs of the timers (S_6, S_7, S_8) and the I/O connector.

Lines RUN and STATUS control the conversion process in A-D converter IC_9. The computer can check for completion of the conversion process by reading STATUS (Ø=conversion complete and data available for reading; 1=conversion not complete). When input RUN is held logic high, conversion cycles are automatically started every 2 ms. When STATUS goes low, the ADC runs an internal autozero procedure, which takes 1.75 μs. The computer can read the conversion result during this interval. Normally, then, the converter needs about 2 ms for completing the conversion and autozero phases. The actual conversion time, however, is shorter, depending on the measured value. The redundant part of the conversion phases can be skipped simply by making RUN low immediately after detecting the low level on STATUS. The converter then enters the autozero phase, and is ready for starting the next conversion when RUN is made high.

The above procedure is illustrated in the flowchart of Fig. 6. The input voltage should be stable during the A-D conversion (STATUS is "1"). The computer can arrange this with the aid of, for instance, a sample and hold circuit controlled by the STATUS signal. The ADC is adjusted with P_1. The voltage at the reference input is set to 2.048 V, so that the "full-scale" input voltage is 4.096 V ($=2U_{ref}$). The data sheets of the ICL7109 should be consulted before setting a reference voltage other than used here.

A-D converter IC_9 obtains its input voltage from 8-channel analogue multiplexer IC_{13}. Input channel selection is effected via Port B in PPI IC_5. The connector for the analogue input voltages also carries three selection signals provided by Port B. These signals are intended for controlling an external analogue output multiplexer.

D-A converter IC_{10} supplies an analogue output current, which converted to a voltage by A_1, while A_2 ensures that the output voltage has an effective range of $-U_{max}$ to $+U_{max}$. The manufacturer of the AD712, Analog Devices, recommends matching of R_7, R_8 and R_9 to within 0.01%. In practice, this can be approximated with acceptable results by using a digital ohm-meter for selecting four resistors with practically equal values from a batch of 20K; 1% types. The actual value is not important. R_7 is then made by connecting two matched resistors in parallel. The consequences of a deviating value of this resistor on the analogue output voltage, U_{an}, are made clear by the equation

$$U_{an} = U_{ref}\left(\frac{n}{4096}\cdot\frac{R_9}{R_7} - \frac{R_9}{R_8}\right) \quad [V]$$

which shows that the ratio R_9/R_8 affects the zero level of the output voltage.

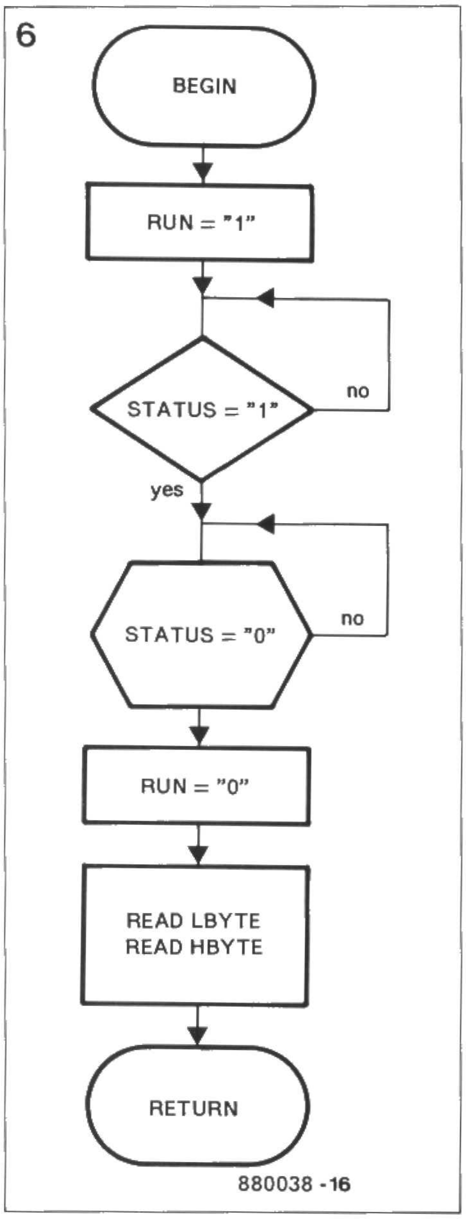


Fig. 6. Suggested flow-diagram for controlling the analogue-digital converter on the I/O extension card.

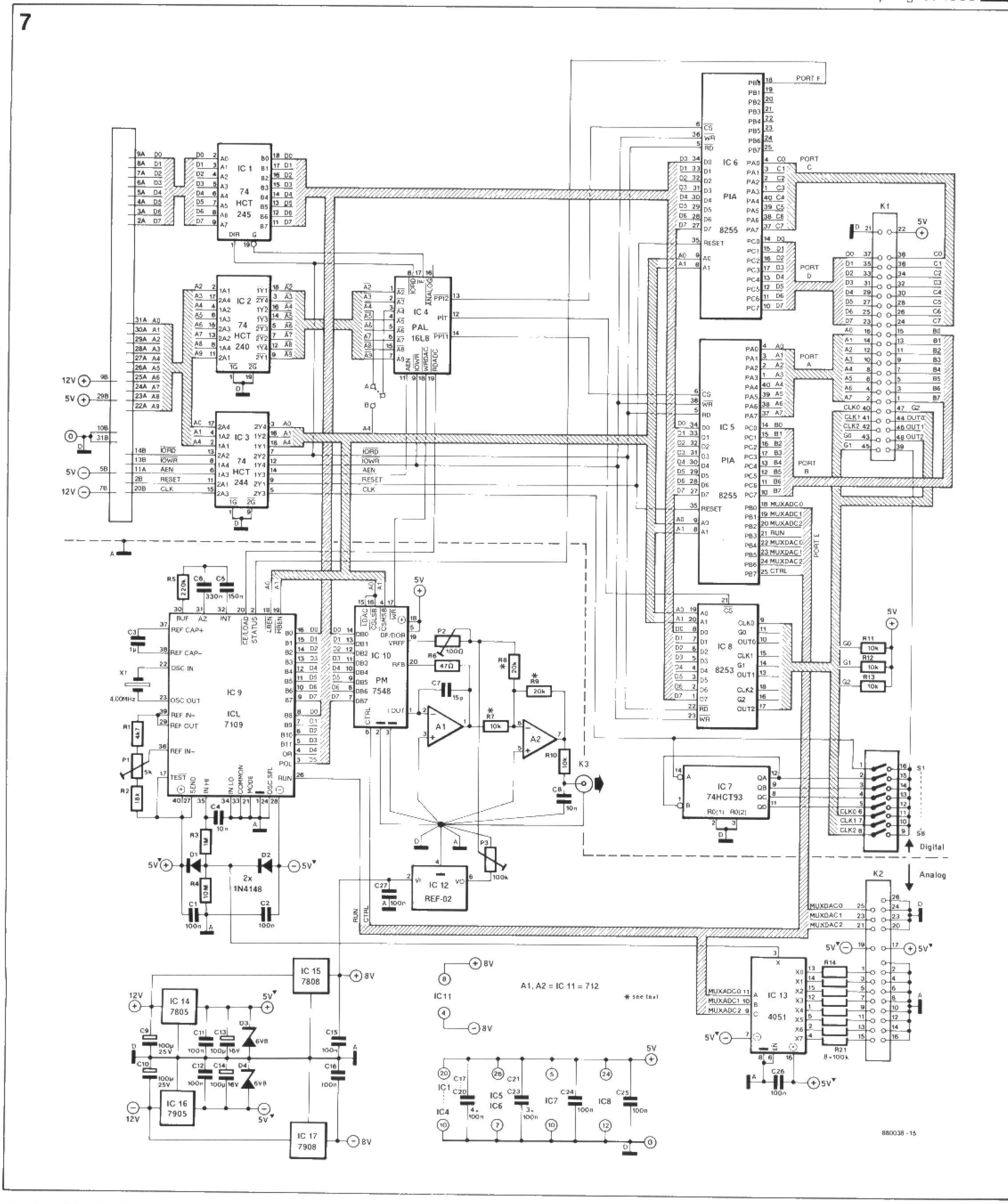


Fig. 7. Circuit diagram of the I/O extension board.

When this ratio deviates from 1, offset becomes impossible to compensate by means of P_2. Ratio R_9/R_7 determines the difference between the maximum and minimum output voltage. For optimum results, this ratio should work out as 2.

The D-A converter obtains its reference voltage from an external source, IC_{12}, because the internal source of ADC IC_9 is unsuitable in this application. The DAC has two adjustments: zero preset P_2, and maximum output voltage preset P_3. An R-C filter at the analogue output removes digital interference from the analogue signal. It should be noted that the R-C time of this filter may have to be slightly reduced when the DAC is operated at "top speed".

The supply on the I/O extension card is perhaps more elaborate than would be expected. The circuit adopted is required, however, to ensure an interference-free supply for the analogue chips on the card. Digital components are fed direct from the 5 V rail provided on the computer bus.

Parts list

Resistors (± 5%):
R_1 = 4K7
R_2 = 18K
R_3 = 1M0
R_4 = 10M
R_5 = 220K
R_6 = 47R
R_7;R_8;R_9 = 4 off matched 20K; 1% resistors. See text.
R_{10}...R_{13} incl. = 10K
R_{14}...R_{21} incl. = 100K
P_1 = 5K0 or 4K7 multiturn preset
P_2 = 100R multiturn preset
P_3 = 100K multiturn preset

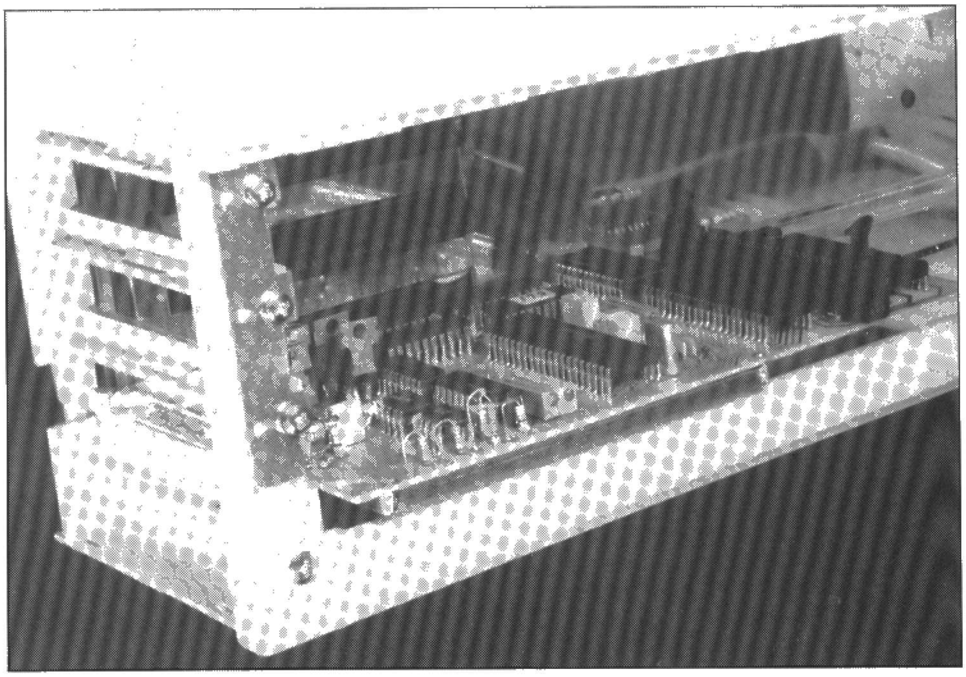
Capacitors:
C_1;C_2;C_{11};C_{12};C_{15}...C_{27} incl. = 100n
C_3 = 1μ0; MKT
C_4;C_8 = 10n
C_5 = 150n
C_6 = 330n
C_7 = 15p
C_9;C_{10} = 100μ; 25 V; radial
C_{13};C_{14} = 100μ; 10 V; radial

Semiconductors:
D_1;D_2 = 1N4148
D_3;D_4 = zener diode 6V8; 400 mW
IC_1 = 74HCT245
IC_2 = 74HCT240
IC_3 = 74HCT244
IC_4 = PAL16L8. Available ready-programmed through the Readers Services under order number ESS561 (unprogrammed chips: ElectroMail order no. 635-527).
IC_5;IC_6 = 8255-5 (Technomatic; Cricklewood)
IC_7 = 74HCT93
IC_8 = 8253-5 (or 8254; see text) (Technomatic; Watford Electronics; Cricklewood).
IC_9 = ICL7109 (Intersil; U.K. distributors are listed on InfoCard 505; *EE* March 1987. Also available from Universal Semiconductor Devices Limited, or C-I Electronics).
IC_{10} = PM-7548 (Precision Monolithics Incorporated. U.K distributors are listed on InfoCard 508; *EE* May 1987. Also available from C-I Electronics).
IC_{11} = OP-215 (PMI) or AD712 (Analog Devices; U.K. distributors are listed on InfoCard 502. Also available from ElectroMail; stock no. 637-810).
IC_{12} = REF-02CP (PMI; C-I Electronics)
IC_{13} = 4051
IC_{14} = 7805
IC_{15} = 7808 IC_{16} = 7905
IC_{17} = 7908

Note: Intel's CMOS 8 MHz versions of the 8253 and 8255 may be used for PCs offering a turbo mode.

Miscellaneous:
S_1...S_8 incl. = 8-way DIL-switch block.
K_1 = 50-way male header with eject handles.
K_2 = 26-way angled header (male) with eject handles.
K_3 = phono socket.
X_1 = quartz crystal 4 MHz.
PCB Type 880038 (see Readers Services page).

This PCB is double-sided, through-plated and provided with gold-plated bus contacts.



Input/output extension card fitted in the rear extension slot of an Amstrad PC1640 SD computer.

9

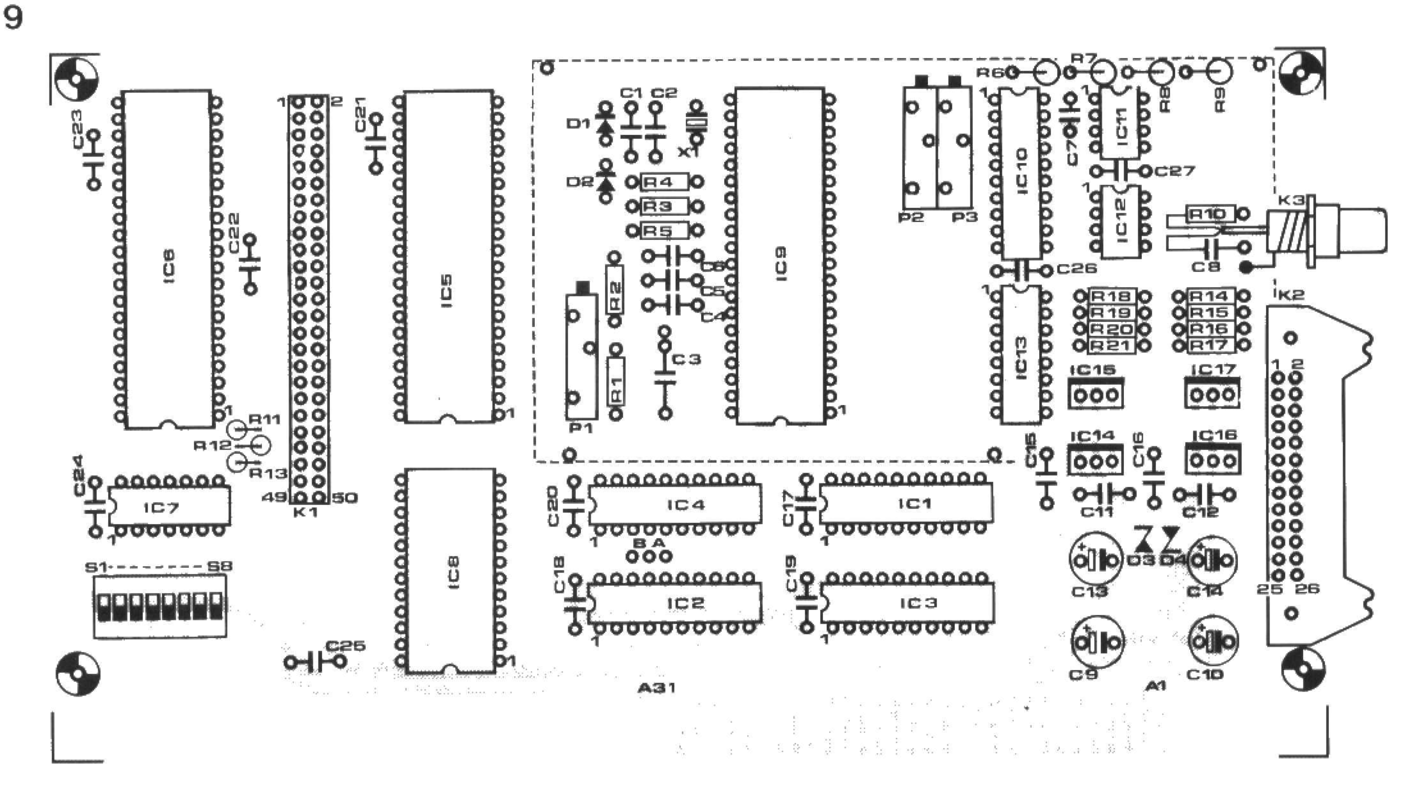


Fig. 9. Component mounting plan of the I/O extension card for PCs and compatibles. Note the screen around the analogue circuitry.

10

```
10 ' ................................................... ibmio interface test
20 X=0: ' ............... address initialisation  0: &H300-&H30F   1: &H310-31F
30 X=&H300+X*&H10
40 ' ................................................... DAC and ADC addresses
50 AH=X+1:  'MS-byte
60 AL=X+2:  'LS-byte
70 A=X+4: E=X+5: B=X+6: C1=X+7:   .............................. I/O addresses
80 C=X+8: F=X+9: D=X+10:C2=X+11
90 T0=X+12:T1=X+13:T2=X+14:C3=X+15
100 DAT =0
110 OUT C1,&H80: ' .................................. E output A and B output
120 OUT C2,&H9B: ' .................................. F input  C and D input
130 OUT E,DAT AND 127: ' ................ left justified data input for the DAC
140  X3=1
150 OUT C3,(0*64+1*16+3*2+1): ' ............................... set counter 0
160 OUT C3,(1*64+1*16+3*2+1): ' ............................... set counter 1
170 OUT C3,(2*64+1*16+3*2+1): ' ............................... set counter 2
180 OUT T0,X3: ' .............................................. start counters
190 OUT T1,X3: ' .................... check timer outputs with an oscilloscope
200 OUT T2,X3
210 ' ............................................................ test I/O ports
220 CLS: LOCATE 23,1:PRINT "Testing I/O"
230 LOCATE 10,1
240 F=0
250 FOR I=0 TO 255
260 OUT A,I: OUT B,I
270 GOSUB 800
280 IF INP (C)<>X THEN F=F+1: PRINT "B out to C in error. Output was: ";I
290 IF F>10 THEN GOTO 310
300 IF INP (D)<>X THEN F=F+1: PRINT "A out to D in error. Output was: ";I
310 IF F>10 THEN I=255
320 NEXT I
330 OUT C1,&H99: ' ............................................ A and B input
340 OUT C2,&H82: ' ........................................... C and D output
350 FOR I=0 TO 255
360 OUT C,I: OUT D,I
370 GOSUB 800
380 IF F>10 THEN GOTO 420
390 IF INP (B)<>X THEN F=F+1: PRINT "C out to B in error. Output was: ";I
400 IF F>10 THEN GOTO 420
410 IF INP (A)<>X THEN F=F+1: PRINT "D out to A in error. Output was: ";I
420 IF F>10 THEN I=255
430 NEXT I
440 OUT C2,&H9B: ' ............................................ C and D input
450 LOCATE 23,1
460 IF F=0 THEN PRINT "I/O ok       " ELSE PRINT "I/O not ok!!!  "
470 LOCATE 1,1: PRINT "ADC INPUT IS:";: X2=0
480 ' ........................................... testing and adjusting ADC
490 ANALOG=0: ' .................................. selected multiplexer input
500 DAT=(ANALOG OR 8) OR (DAT AND 247): ' ......................... set run bit
510 OUT E,DAT
520 IF INP(F) AND 1 =0 THEN 520
530 IF INP(F) AND 1 =1 THEN 530
540 DAT=ANALOG OR (DAT AND 247): ' ............................ reset run bit
550 OUT E,DAT
560 X1= ((INP(AH) AND 48)) : X2= (((INP(AH) AND 15) *256 + INP(AL)))
570 IF (X1 AND 16) = 16 THEN PRINT "overflow     ": GOTO 610
580 IF (X1 AND 32) =0  THEN X2= -X2

                                   1

590 PRINT "  ";:PRINT USING "#####.#";((X2 * 2000!)/ &H800) ; :PRINT"mV  "
600 LOCATE 7,1: PRINT "ADJUST P1 UNTIL READING IS IN ACCORDANCE WITH MULTIMETER"
610 LOCATE 9,1: PRINT "PRESS SPACE TO CHECK AND ADJUST DAC                 "
620 IF INKEY$<>" " THEN 470
630 ' ........................................... testing and adjusting DAC
640 LOCATE 7,1: PRINT "                                                        "
650 OUT E,DAT AND 127
660 OUT AH,&H80: OUT AL,0
670 LOCATE 1,1: PRINT "ADJUST P2 FOR DAC OUTPUT OF 0.0 mV              "
680 LOCATE 9,1: PRINT "PRESS SPACE FOR THE NEXT DAC CHECK"
690 IF INKEY$<>" " THEN 690
700 OUT AH,&HFF: OUT AL,&HF0
710 LOCATE 1,1: PRINT "ADJUST P3 FOR DAC OUTPUT OF 3.998 V             "
720 LOCATE 9,1: PRINT "PRESS SPACE FOR THE NEXT DAC CHECK"
730 IF INKEY$<>" " THEN 730
740 OUT AH,&H0: OUT AL,&H0
750 LOCATE 1,1: PRINT "THE OUTPUT OF THE DAC SHOULD NOW READ -4.000 V"
760 LOCATE 9,1: PRINT "PRESS SPACE TO END PROGRAM          "
770 A$=INKEY$: IF A$="" THEN 770
780 IF A$<>" " THEN 650
790 CLS: END
800 ' ................................................... reorganizing bits
810 X=0
820 FOR J=0 TO 7
830 IF (I AND (2^J))<> 0 THEN X=2^(7-J)+X
840 NEXT
850 RETURN
```

880038 -18

Fig. 10. Use this GWBASIC program to test the card for correct operation.

Construction

The I/O card is relatively simple to build, provided the soldering work is done with care and precision. It is recommended to use sockets with turned pins for all ICs. The double-sided, through-plated PCB supplied through the Readers Services (Fig. 9) is a type with gold-plated bus contacts as usual for extension IBM extension cards.

The analogue circuit section is screened as shown by the dashed lines on the component overlay. This requires bending voltage regulators IC_{14}...IC_{17} incl. towards R_{17} and R_{21}. Beware of short-circuits between metal tabs of IC_{15}; IC_{17} and the screen and/or the resistor terminals. Components R_{10} and C_8 should be fitted as close as possible to phono socket K_3 as shown in the photograph of Fig. 11.

Table 5.

Address decoder for PC I/O card

```
device 16L8

/A2        1
/A3        2
/A4        3
/A5        4
/A6        5
/A7        6
/A9        7
/IORD      8
/IOWR      9
GND        10
AEN        11
/PIT       12
/PPI2      13
/PPI1      14
/A8        15
/IO0       16
/E         17
/WRDAC     18
/RDADC     19
VCC        20

MACRO F0 /A4*/A5*/A6*/A7*A8*A9*/AEN;
MACRO F1 /A2*/A3*E;

START
/IO0       /=&F1;
/RDADC     /=IORD*&F1;
/WRDAC     /=IOWR*&F1;
/PPI1      /=A2*/A3*E;
/PPI2      /=/A2*A3*E;
/PIT       /=A2*A3*E;
/E         /=&F0*IORD+&F0*IOWR;
END
```

Table 5. Programming data for the PAL, which combines the functions of a number of SSI logic gates.

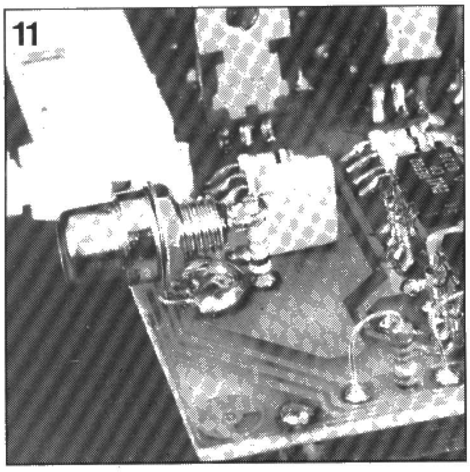


Fig. 11. The filter components at the analogue output of the card are soldered direct to the centre pin of phono socket K_3.

Mapping the card

The I/O extension can be addressed in block $30x_H$ (jumper A) or $31x_H$ (jumper B). Table 6 shows the address assignment in the I/O block.

The PPIs should be initialized by writing $100xx00x_2$ in the control register of IC_5, and $100xx01x_2$ in that of IC_6. Bits x are programmed to requirement. Port E is always used in the output mode, Port F in the input mode. The other ports can be set to input or output.

Table 6.

I/O card register overview:

Address (hex)	Read	Write
3x0	not allowed	not allowed
3x1	MS byte of ADC	MS byte of DAC
3x2	LS byte of ADC	LS byte of DAC, load input register in DAC register
3x3	not allowed	not allowed
3x4	Port A	Port A
3x5	Port E	Port E
3x6	Port B	Port B
3x7	not allowed	Control register IC5
3x8	Port C	Port C
3x9	Port F	Port F
3xA	Port D	Port D
3xB	not allowed	Control register IC6
3xC	Timer Ø	Timer Ø
3xD	Timer 1	Timer 1
3xE	Timer 2	Timer 2
3xF	not allowed (but o.k. for 8254)	Control register IC8

Note: x = 0 or 1 depending on address allocation (jumper on card).

Adjustment and testing

In view of the 12-bit accuracy required, a good quality 3¾-digit digital multimeter is essential for adjusting the ADC and DAC. The former is adjusted by applying 4 V to it and having the computer print the conversion result on screen with the aid of a programmed loop. Adjust P_1 until the decimal value of the conversion results corresponds to the voltage read by the DMM.

The DAC is adjusted as follows: set the DMM to the appropriate voltage range, and connect it to the analogue output, K_3. To begin with, the zero level must be set. Write 800_H to the DAC, and adjust P_2 for an output voltage of nought. Next, write FFF_H, and adjust P_3 for the desired maximum value.

The I/O extension card can be tested with the aid of the GWBASIC program listed in Fig. 10. Use 10 kΩ resistors for connecting Port A to Port D, and Port B to Port C. The bit-order is reversed in these connections, i.e., bit Ø of Port A goes to bit 7 of Port D, bit 1 to bit 6, etc. *R*

Reference:
Peripheral Design Handbook. Intel, 1981.

PEOPLE

Mr **Tony Ventrella**, Legal Adviser to BEAMA, has taken up the additional appointment of Company Secretary. He takes over from Mr **Derek Shrivell**, Company Secretary and Financial Controller, who is retiring after 18 years' service with BEAMA.

The financial duties will be undertaken by Mr **Richard Leishamn**, who has been BEAMA's Chief Accountant for six years.

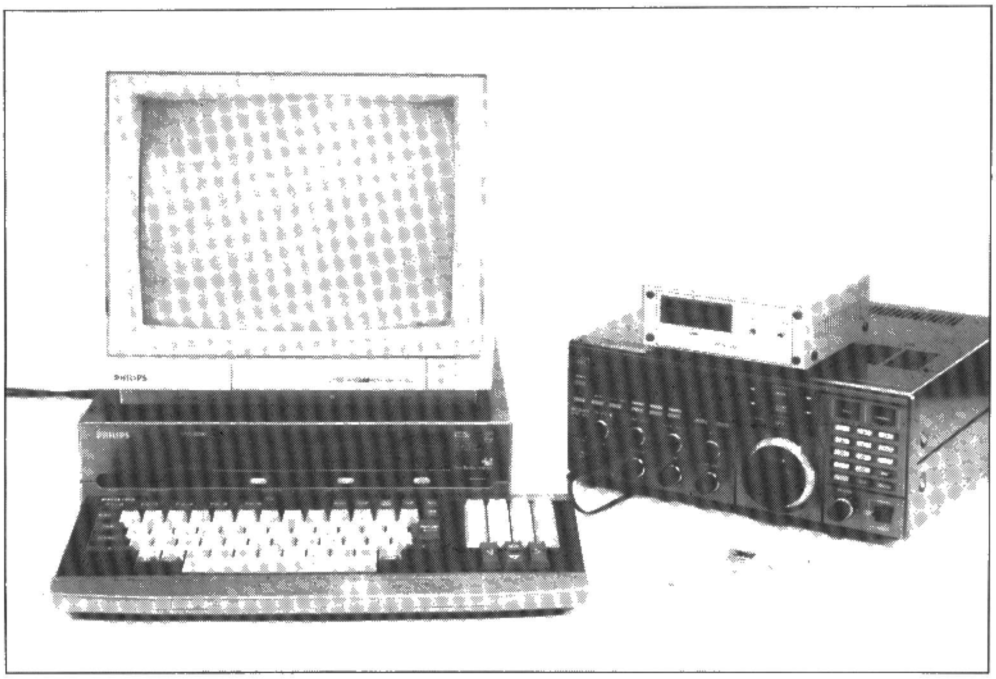
Mr **Derek P. Hornby** has been appointed Chairman of the National Accreditation Council for Certification Bodies (NACCB) in succession to Professor John Ashworth, who is retiring at the end of his three-year period of office. The appointment carries with it ex-officio membership of the BSI board.

Mr **Heinz Süsse** has been appointed as European Technical Director for CORCOM, the well-known manufacturer of mains interference suppression filters. Based at CORCOM HQ in Munich, he will also be responsible for applications support in the UK via the CORCOM operation in East Kilbride.

Siliconix has appointed **Brian Wadsworth** as European Distributor Marketing Manager, responsible for the company's network in the UK and Europe. Mr Wadsworth will be based at Siliconix' European HQ at Newbury, Berkshire.

Cox Associates Ltd (CAL) has announced the appointment of Alan Luxton as Production Manager with responsibility for all the company's manufacturing activities.

RECEPTION AND TRANSMISSION OF RADIOTELETYPE



The results of our recent Readership Survey have confirmed the popularity of construction projects and informative articles on every aspect of modern telecommunications. In particular, RTTY is a well-liked subject; and for good reasons, because the combination of simple circuits and a computer offers many advantages over the mechanical telex machine. In addition to a low-cost telex converter plus tuning aid and powerful decoding software, we present a precision, phase-synchronous, AFSK generator for radio amateurs in possession of a licence to transmit RTTY.

If you have never worked with telex for fear of complex circuits, this is the time to take the decisive step into the fascinating world of long-distance communication where messages originate from press bureaus, radio amateurs, navigational, meteorological and utility stations in the shortwave bands. In this article it is assumed that the reader is familiar with the concept of RTTY, but just for convenience a short recapitulation is offered on the structure of the serial data format.

Baudot code and FSK

RTTY information is composed of characters that are transmitted sequentially to the format shown in Fig. 1. One dataword is composed of 7 bits:

- 1 start bit, which is always low;
- 5 data bits, which represent the character;
- 1 stop bit, which has a duration 1.5T rather than T used for the start and data bits.

Five databits offer 32 (2^5) possible combinations. There are, however, 26 letters in the alphabet, 10 numbers, and a number of punctuation marks and symbols, so that 32 positions would appear too few. A solution to this problem is offered by the Baudot system, which reserves a special code for distinguishing between letters on the one hand and figures, punctuation marks or signs on the other. This means that a number of available databit combinations can represent either a letter or a number, depending on the group selection code, which is transmitted after pressing the *Ltrs* or *Figs* key

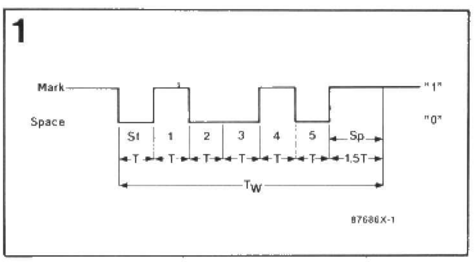


Fig. 1. Serial transmission of an RTTY character.

on the teletype machine (*cp* the caps lock function on a typewriter).

The bit combinations in the Baudot system are listed in Table 1 for reference. It should be noted that there exist many other systems and standards for RTTY, with variations in transmission formats, pulse duration and bit combinations. The majority of telex stations, however, use the Baudot code.

The speed of the RTTY transmission is called the *baud rate*. This is defined as the number of bits transmitted per second. For instance, a 75 baud station can transmit nearly 11 characters per second.

On the shortwave bands, the modulation used for RTTY transmission is usually frequency shift keying (FSK) — depending on the logic level of the bit (mark or space), frequency f_1 or f_2 is transmitted. The difference between these frequencies is called *shift*.

Figure 2 summarizes the above by showing the units required for over-air transmission of a telex message. This article will deal with the construction of the RTTY converter, AFSK generator and

2

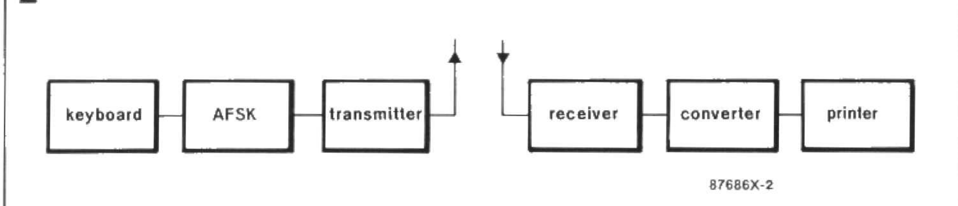


Fig. 2. Transmission and reception of radioteletype.

printer. The latter function is assumed by a computer running an advanced program that enables decoding RTTY at a number of standardized baud rates.

PLL-based RTTY converter and tuning aid

The function of the RTTY converter is to translate the mark and space notes received into the appropriate logic levels. To keep the circuit as simple as possible without compromising versatily, it was decided to set it up around an integrated phase-locked loop, the Type XR2211 from Exar.

3

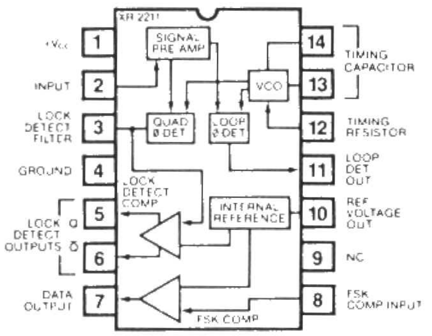


Fig. 3. Pinning of the Type XR2211 phase-locked loop from Exar.

The operation of the converter is best explained with reference to the internal circuit of the XR2211 shown in Fig. 4, and the circuit diagram in Fig. 5. Note that the small circles and numbers in Fig. 4 are the pins of the chip — the resistors and capacitors are external components. Basically, a phase-locked loop will attempt to make the frequency of the internal voltage-controlled oscillator (VCO) equal to that of the input signal. The two frequencies are compared by the quad and/or loop detector, and the error signal produced by one of these is used for controlling (i.e., tuning) the VCO until the difference between VCO frequency and input frquency is nought. The VCO control signal is, therefore, a measure of the frequency of the input signal provided by the shortwave receiver tuned to an RTTY transmission. In the present application of the XR2211, the on-board quad detector is not used.

The circuit diagram of the converter is very similar to the block diagram of the XR2211. The input signal enters the IC via R-C network R_1-C_1-C_2, and is fed to an amplifier driving the loop detector, whose output signal is used for controlling the VCO via potential divider R_7-R_8, and driving comparator 1 via R_6-C_4. The comparator uses the input voltage to decide whether the received frequency is f_1 or f_2 (mark or space). Its output signal needs filtering, however, because the PLL is so fast that it also responds to interference in the input signal. Comparator 2, in combination with low-pass R_2-C_3, ensures a filtered, rectangular output signal. The amplitude of the output signal of comparator 1 (fed to pin 3 via R_2-C_3) is much greater than that supplied by the quad detector, which is thus rendered ineffective. Comparator 2 has complementary outputs Q and $\overline{Q}$, which are useful for selecting between stations transmitting standard and inverted RTTY signals (logic high = f_0; logic low = f_1). This selection is accomplished by toggle switch S_1.

It will be clear that the mark and space tones can only be decoded correctly when the receiver is accurately tuned to the telex transmitter. Figure 6 shows the circuit diagram of a bar-graph tuning indicator which will prove indispensable for RTTY reception. Based on a VU-meter IC, it indicates centre tuning as well as shift employed by the transmitter. The display indicates frequency of the mark and space tones by means of LEDs to the right and left of the centre, respectively. The centre two LEDs are illuminated when the input signal from the converter has a frequency of 1500 Hz. When the receiver is correctly tuned, LEDs at equal distance from the centre will be seen to flash rapidly. The distance between the illuminated LEDs is a measure of the shift employed by the RTTY transmitter.

Table 1

bit 1 2 3 4 5	letters	figures
1 1 0 0 0	A	—
1 0 0 1 1	B	?
0 1 1 1 0	C	:
1 0 0 1 0	D	who are you
1 0 0 0 0	E	3
1 0 1 1 0	F	optional
0 1 0 1 1	G	optional
0 0 1 0 1	H	optional
0 1 1 0 0	I	8
1 1 0 1 0	J	bell
1 1 1 1 0	K	(
0 1 0 0 1	L	)
0 0 1 1 1	M	.
0 0 1 1 0	N	,
0 0 0 1 1	O	9
0 1 1 0 1	P	0
1 1 1 0 1	Q	1
0 1 0 1 0	R	4
1 0 1 0 0	S	'
0 0 0 0 1	T	5
1 1 1 0 0	U	7
0 1 1 1 1	V	=
1 1 0 0 1	W	2
1 0 1 1 1	X	/
1 0 1 0 1	Y	6
1 0 0 0 1	Z	+
0 0 0 1 0	carriage return	
0 1 0 0 0	line feed	
1 1 1 1 1	letters	
1 1 0 1 1	figures	
0 0 0 0 0	space	
0 0 0 0 0	backspace	

4

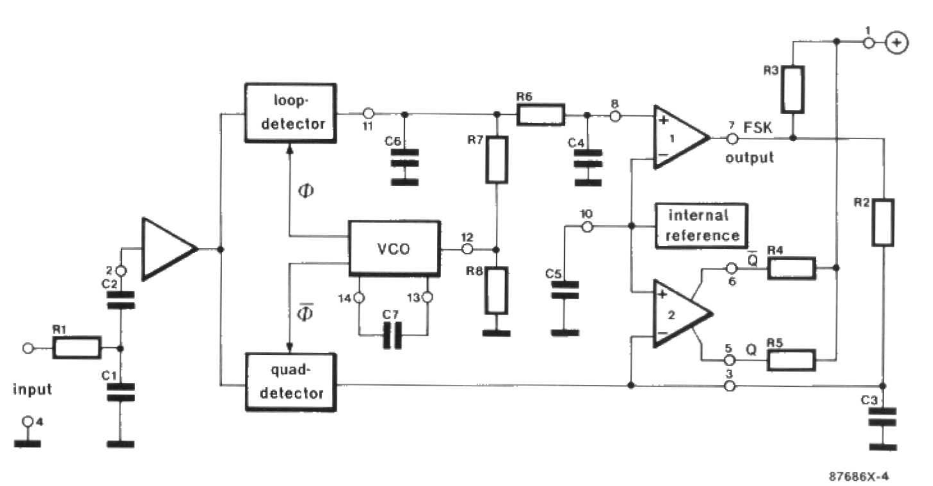


Fig. 4. Internal circuit of the XR2211.

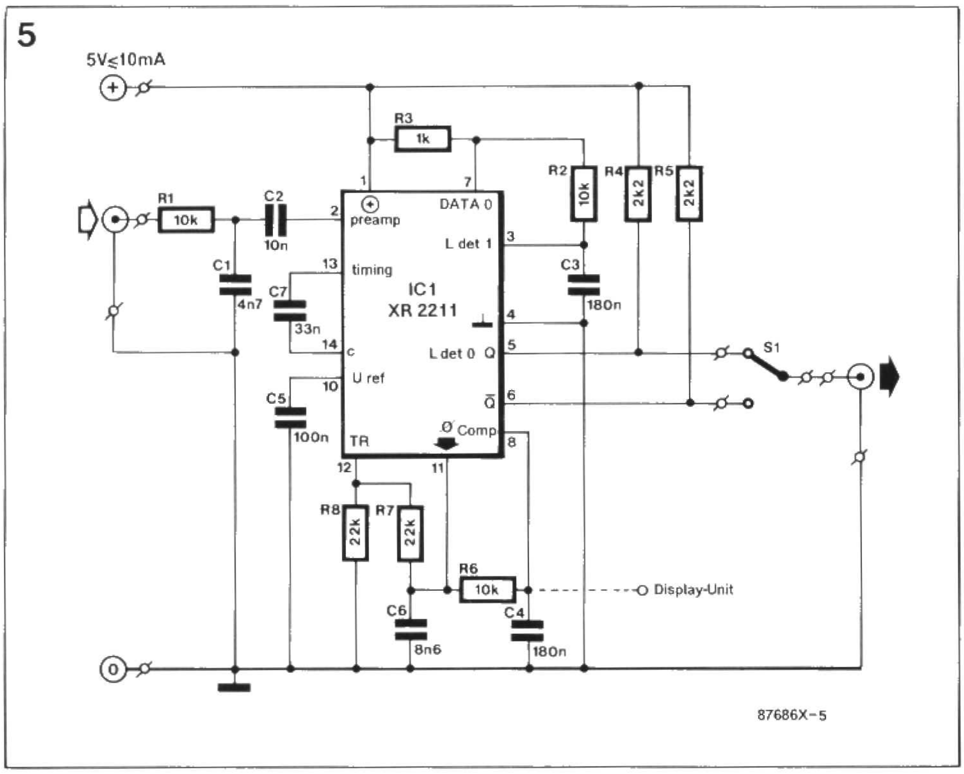


Fig. 5. Circuit diagram of the RTTY decoder.

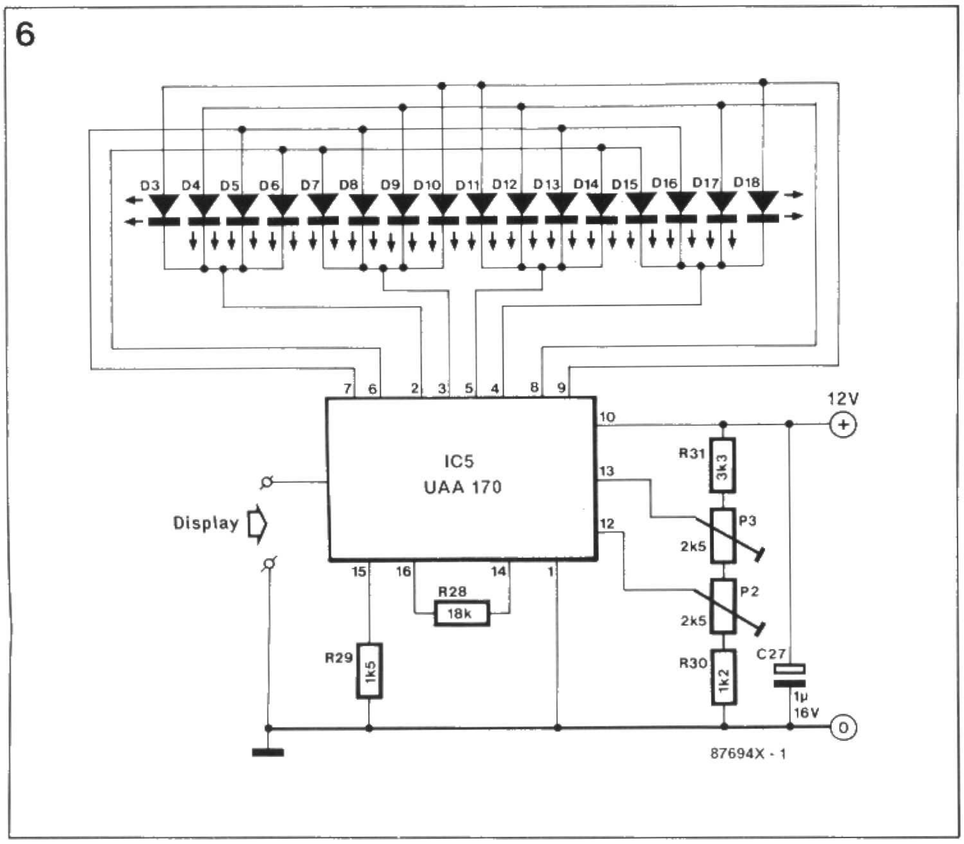


Fig. 6. Circuit diagram of the optional RTTY tuning aid, which is essentially a LED-based shift and tuning indicator.

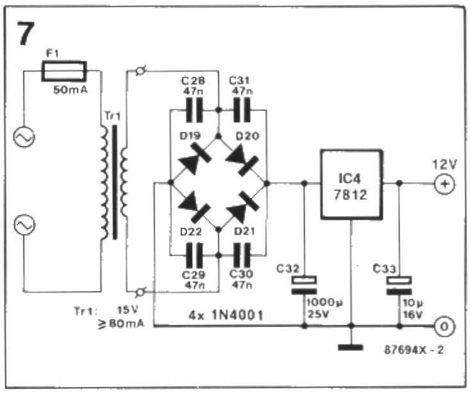


Fig. 7. Suggested power supply for the RTTY converter and tuning aid.

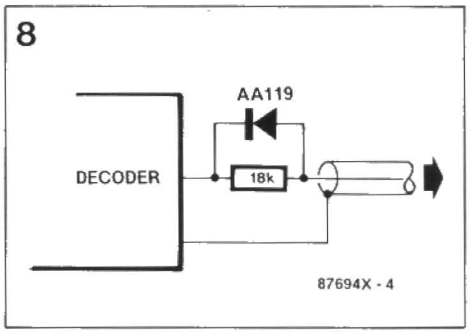


Fig. 8. Resistor-diode attenuator for reducing the logic swing at the output of the decoder from 12 V_{pp} to 5 V_{pp}.

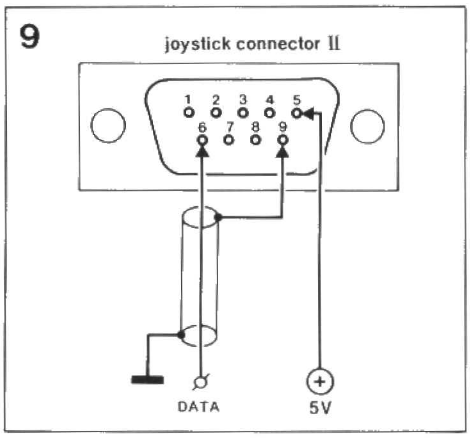


Fig. 9. Connection of the RTTY decoder to joystick connector #2 on an MSX 1 or MSX2 computer.

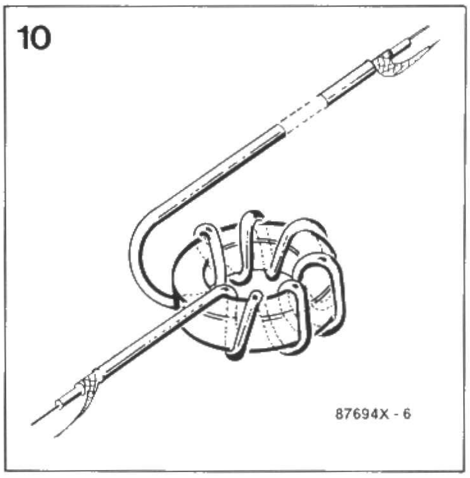


Fig. 10. Interference from the computer can be suppressed by winding the serial data and supply cables onto a ferrite ring core.

Power supply and computer interface

It will be noted that the supply voltage for the RTTY converter is 5 V, whereas that for the display unit is 12 V. The latter is optional, so that the converter may be fed from the computer's built-in 5 V power supply. When the display is used in conjunction with the converter and a computer, it is essential that the converter is fed from 12 V rather than 5 V. This can be done with impunity using the power supply of Fig. 7 and the simple computer interface of Fig. 8 (the latter reduces the converter's output voltage swing of 12 V_{pp} to 5 V_{pp}).
The serial bitstream from the converter is applied to the computer's joystick port as shown in Fig. 9. To prevent

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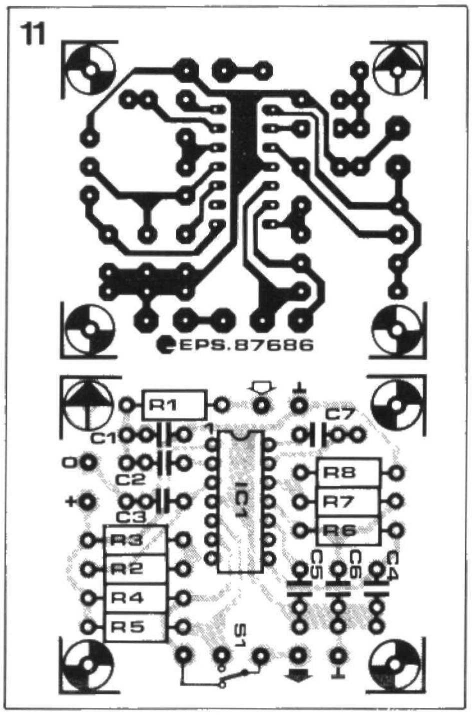


Fig. 11. Printed circuit board for the single-chip RTTY decoder.

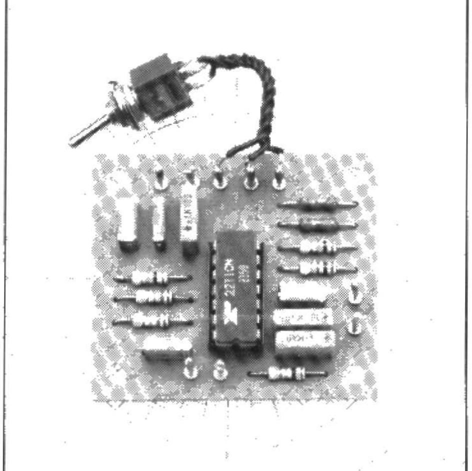
Parts list
RTTY DECODER. CIRCUIT DIAGRAM: FIG. 5.

Resistors (±5%):
R1;R2;R6 = 10K
R3 = 1K0
R4;R5 = 2K2
R7;R8 = 22K

Capacitors:
C1 = 4n7
C2 = 10n
C3;C4 = 180n
C5 = 100n
C6 = 8n2
C7 = 33n

Semiconductor:
IC1 = XR2211 (Manufacturer: Exar. Listed by Maplin, Cricklewood Electronics, Universal Semiconductor Devices)

Miscellaneous:
S1 = miniature SPDT switch.
PCB Type 87686X (not available through the Readers Services).



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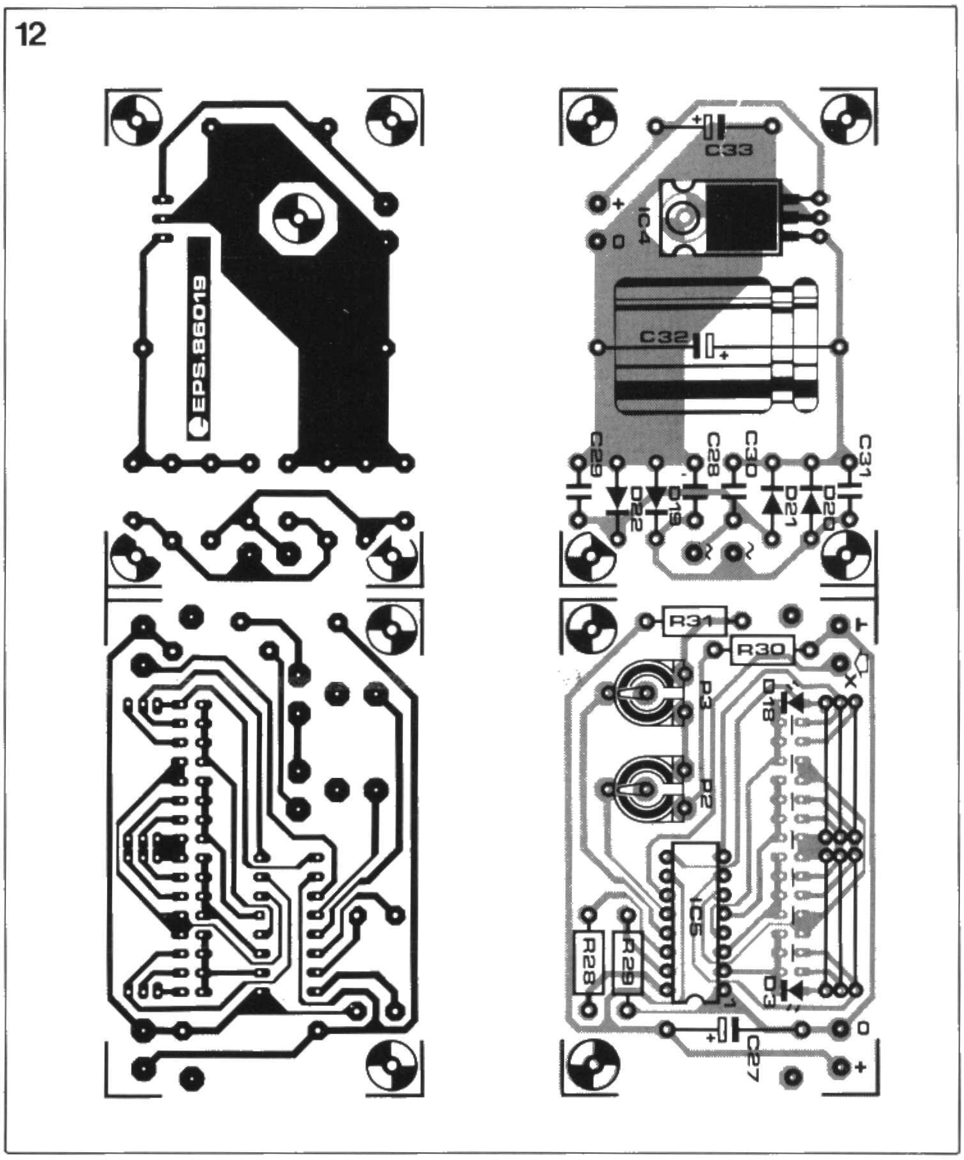

Fig. 12. Printed circuit board for the RTTY tuning aid and power supply.

digital interference from the computer entering the converter circuit, it is recommended to wind the cable between the converter and the computer onto a medium-size ferrite ring core as shown in Fig. 10.

Setting up

Apply the output signal of a sine wave generator to the converter. Adjust the generator between 1 and 2 kHz, and use a scope or a voltmeter to find the frequency at which the output of the converter toggles. Leave the generator set to this frequency, and adjust P_2 and P_3 in the display circuit until the two centre LEDs light. Then correct the settings until 1 kHz corresponds to the leftmost LED, and 2 kHz to the right-most LED. The preset adjustments will be found to interact, and care should be taken not to shift the centre frequency indication away from D_{10}-D_{11}. If necessary, repeat the adjustments.

When a sine wave generator is not available, connect the converter to the receiver, and tune to a station that transmits an unmodulated carrier.

Parts list
TUNING AID AND POWER SUPPLY.
CIRCUIT DIAGRAMS: FIGS. 6 & 7.

Resistors (±5%):
R28 = 18K
R29 = 1K5
R30 = 1K2
R31 = 3K3
P2;P3 = 2K5 or 2K2 preset H

Capacitors:
C27 = 1µ0; 16 V
C28 . . . C31 incl. = 47n
C32 = 1000µ; 25 V
C33 = 10µ; 16 V

Semiconductors:
D3 . . . D18 incl. = rectangular LED
D19 . . . D22 incl. = 1N4001
IC4 = 7812
IC5 = UAA170 (listed by Maplin, Cricklewood)

Miscellaneous:
F1 = 250 mA fuse, delayed action, with chassis-mount holder.
Tr1 = mains transformer, 15 V; ≥80 mA.
PCB 86019 (not available through the Readers Services).

Switch to SSB, and tune the receiver until the output of the converter toggles. The note produced by the receiver then has a frequency of about 1500 Hz, and P_2-P_3 may be adjusted such that the centre LEDs light. Table 2 lists a number of meteorological services that use a shift of 1 kHz, enabling the maximum shift indication of the display to be set as discussed above.

Table 2

speed	frequency (kHz)	service
50	3038	Meteo
50	13530	Meteo
75	2474	Marine
75	4280	Marine
100	7980	Meteo

Enter the computer

An RTTY decoder program was written for MSX1 and MSX2 computers with a disk drive running under MSXDOS. This program can be obtained through the Readers Services by sending in a formatted 3½-inch diskette that contains MSXDOS.SYS and COMMAND.COM. The program has its own video drivers because those available under standard MSX software proved too slow to ensure correct interrupt servicing and picture scrolling given the speed of the pulses supplied by the RTTY converter. Further, the program boots up automatically when the disk is inserted in the machine at power-on. After some time, the main menu will appear, and the user is prompted to enter the required parameters — see Fig. 13.

Table 3

speed	frequency (kHz)	country	time (approx., GMT)
100	16046	USSR	9.00 – 13.00 h.
50	15633	N. Korea	12.00 – 13.00 h.
50	13780	N. Korea	12.00 – 13.00 h.
50	13524	Iraq	13.00 – 17.00 h.
75	13770	USA	13.00 – 17.00 h.
50	7960	Iran	19.00 – 24.00 h.
75	5460	USA	0.30 – 3.30 h.
75	6941	USA	0.30 – 3.30 h.

Although the menu is mostly self-explanatory, a short discussion will be given of the various options available.

- **ASCI1:** when this mode is selected, the computer switches to decoding 8 databits and 2 stop bits. Since this is a non-standard format, the mode is called ASCI1 rather than ASCII. It can be tried by tuning to the news bulletin of radio amateur station W1AW on 14.095 MHz in the 20 m band. This North-American station transmits at 110 baud, and is one of very few not to use the Baudot code.
- **Shift:** It was already noted that the Baudot code has two special characters for switching between letters (*Ltrs*) and figures/punctuation marks/signs (*Figs*). When SHIFT NORMAL is selected from the menu, switching between these character sets takes place when the relevant code is received. When, owing to interference, the computer does not receive one of these codes, it will produce illegible text. To prevent this, select Unshift On Space (UOS), to automatically switch to letters following the reception of a space character. This is particularly useful for stations that transmit mainly text, e.g. press bureaus. The reverse is also possible: Shift On Space (SOS) causes the 'figures' set to be selected following a space, aiding in the reception of, for instance, meteorological data, which is mainly clusters of figures and signs.
- **Invert data:** this is the software equivalent of operating the polarity switch, S_1, on the RTTY converter.
- **Baud rate selection:** available speeds are arranged in order of frequency of use.
- **Read Text buffer:** text read by the computer is stored in memory. When the computer is in the RX (receive) mode, the memory contents can be examined and, if required, sent to a printer. The text buffer mode is selected by pressing M on the keyboard. Decoding of incoming data is inhibited in this mode, which can be left by pressing R.

User-selected parameters are loaded by pressing the space bar to enter the receive (RX) mode. When the settings are correct, and the converter is properly aligned, text will appear on the screen. If there appears no text, or only garbled data, the parameters must be changed by returning to the menu — press RETURN.

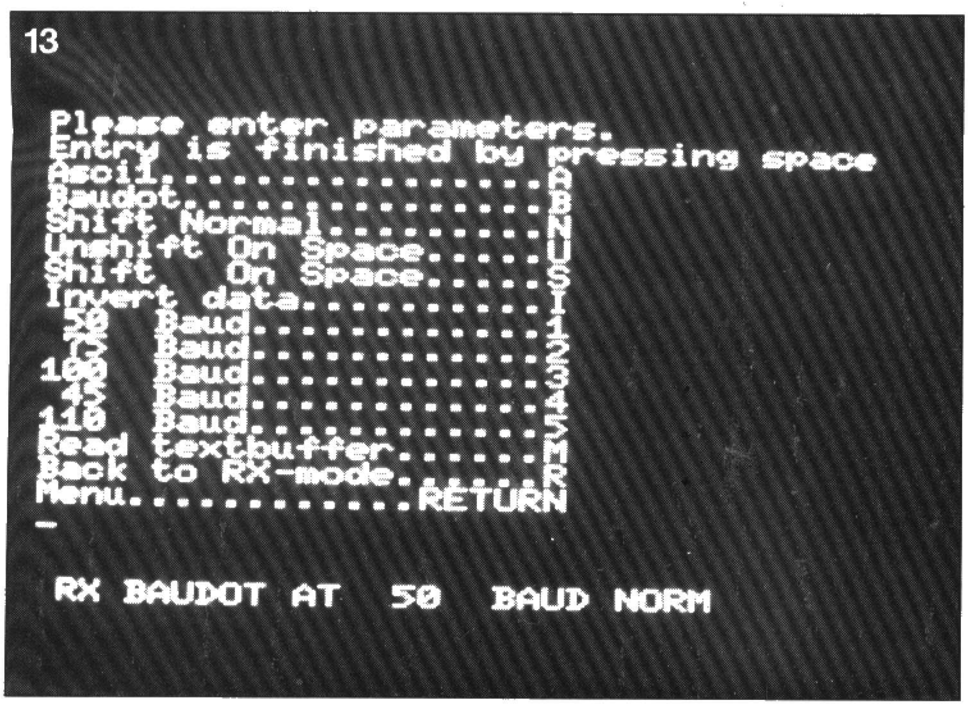

Fig. 13. Menu screen shown by the RTTY decoding program for MSX computers.

An interesting feature of the MSX RTTY software is its ability to decode Russian stations using an extended version of the Baudot code. A number of additional shift characters are used in this to enable using the Cyrillic alphabet, which has more than 26 letters. The software translates these in characters that are relatively little used on computers. For instance, the ampersand sign, &, represents the sound /sj/ (Sa&a), while the exclamation mark, !, stands for /è/ (!lektrik). There are more replacement signs, but a discussion of all of these is beyond the scope of this article. Actually receiving Russian stations is the best way of getting accustomed to the special signs, since their sound and meaning can be learned fairly rapidly by deduction from the context. Recommended frequencies are around 8,350 kHz (evenings and night) and 12,500 kHz (daytime). Other useful frequency allocations are given in Table 3.

Finally, it will be understood that usable results are only obtained by using a good-quality SSB receiver connected to an outdoor aerial. Everything possible must be done to eliminate sources of interference, since these give rise to decoding errors even when the telex station received has sufficient field-strength locally.

The RTTY decoding program for MSX computers can be obtained through the Readers Services under number **XSS100** by sending a formatted 3½-inch diskette containing MSXDOS.COM and COMMAND.COM to our Brentford office. The cost of programming this diskette is stated on the Readers Services page in this issue. A stamped, self-adressed envelope must be included for return postage (overseas readers: please include 5 IRCs). Listings or tape versions of the program are not available.

AFSK generator for VHF RTTY

Contrary to a good many other designs, the audio frequency shift keying (AFSK) generator described here is not only quartz-controlled, but also phase-synchronous, ensuring very high signal quality and accurate definition of the tone frequencies. Most AFSK generators used by radio amateurs incorporate R-C or L-C oscillators, which often suffer from poor stability and accuracy. Many PLL-based and computer-driven AFSK generators are also remarkable for their low signal quality and deviation from the prescribed mark and space frequencies, making it impossible to achieve proper decoding when the receiving station uses filters that are known to be accurately dimensioned to the relevant RTTY standard. As a result, a poor signal-to-noise ratio is obtained, and there are station operators who attempt to cure this by switching over to AMTOR or other error-correcting systems, while the trouble originates clearly from their poorly designed AFSK generator.

The circuit described here is simple to build from a handful of fairly common parts, and can produce virtually any frequency used for AFSK at a resolution of a few hertz.

AFSK generator: principle of operation

The block diagram of Fig. 14 shows that a quartz-controlled 10 MHz oscillator clocks two counters/dividers, whose divide ratio can be preset to any value between 1 and 256. The counters are, of course, configured to give signals of different output frequency, which are applied to an electronic toggle switch composed of NAND gates. The position of this switch is controlled by the logic level of the signal applied to the TTL input. The frequency of the signal at the output of the electronic switch changes with every change in the logic level of the signal at the digital input. In this arrangement, phase disturbance will occur because of the abrupt switching between the output frequencies. The result is an annoying click, which is particularly troublesome when SSB is used, because it easily causes sideband splatter. These switching problems are also frequently encountered in PLL-based AFSK generators, where the oscillator is constantly on the verge of losing lock owing to the fast changing AFSK frequency.

The present generator offers an elegant way round the above difficulties. The output signal of the electronic toggle switch is not fed direct to the circuit output, but is first applied to a :64 divider. When, owing to the switching at its input, the :64 divider receives one pulse too many or too few, the toggling of the output will be delayed or speeded up by only 1/64th part of the period of the signal applied. This results in negligible pulse-width distortion whilst ensuring that the output signals of the divider are phase-synchronous. After filtering in a low-pass, an RTTY signal is obtained that is much cleaner than one produced by a PLL-based generator.

AFSK generator: circuit description.

The practical circuit diagram is shown in Fig. 15. The 10 MHz clock oscillator is set up around T_1, which drives counters IC_1 and IC_2. The divisor of each of these can be set by means of $S_{1a}\ldots S_{1h}$ and $S_{2a}\ldots S_{2h}$ respectively. The user is left free to use either (DIP-) switches or wire links in these positions. The previously mentioned NAND gates that form the electronic toggle switch are found in IC_3, while IC_4 is the :64 divider. The circuit around T_2 and T_3 forms an interface between the TTL input and IC_3. TTL signals are limited by D_1 and inverted by T_2. T_3 once more inverts the signal. Switch S_3 allows selecting normal or reverse signal polarity.

Harmonics and interference at the output of divider IC_4 are eliminated in low-pass filter L_1-L_2-C_4-C_5-C_6-C_8. The amplitude of the AFSK signal can be set with P_1.

Construction and use

The AFSK generator is relatively easy to construct on Veroboard. In practice, it will be found that a number of configuration switches may be omitted because a fixed set of AFSK frequencies often suffices.

Standard AFSK frequencies and associated shifts are listed in Fig. 16. It is seen that the step size is 85 Hz. Formerly very popular frequencies were 2125 Hz and 2975 Hz. These multiples of 425 Hz, and those of 170 Hz, were long used by American telegraph companies. Dividing these frequencies reveals that they are all multiples of 17 Hz, which is important to remember for accurately setting the AFSK generator. The need for the new AFSK frequencies in Fig. 16 arose from the fact that amateur radio receivers incorporated AF filters too narrow to pass 2975 Hz, so that many RTTY hams were forced to use lower frequencies.

It should be noted that some stations use shifts other than those shown, e.g. 85 Hz or even shifts which are not a multiple of 17 Hz (70 and 240 Hz). When the AFSK generator is used with an SSB transmitter, the absolute frequencies are not important, only the shift. For FM

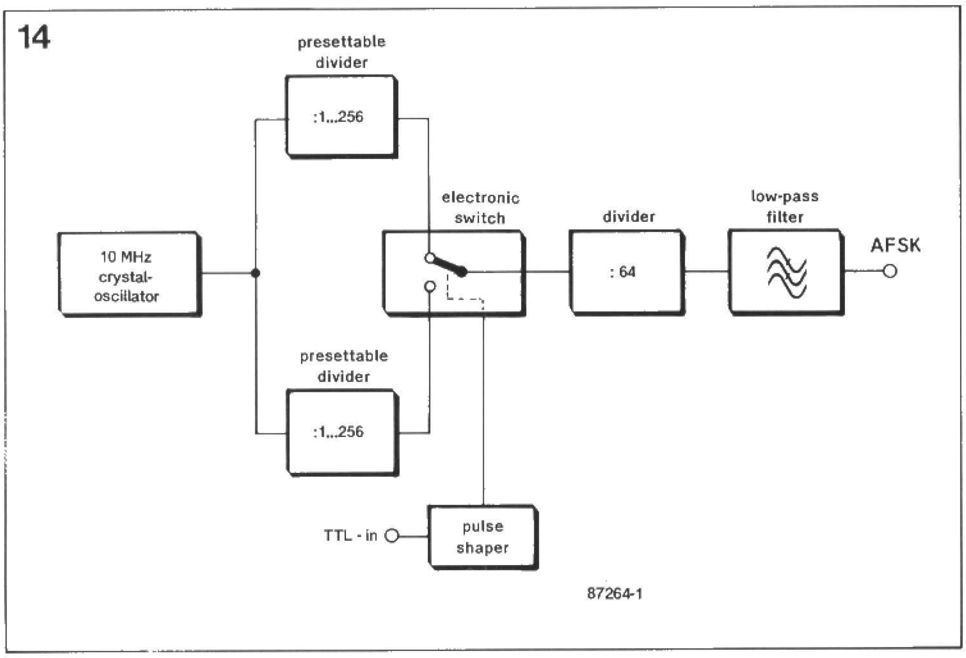

Fig. 14. Block diagram of the crystal-controlled AFSK generator.

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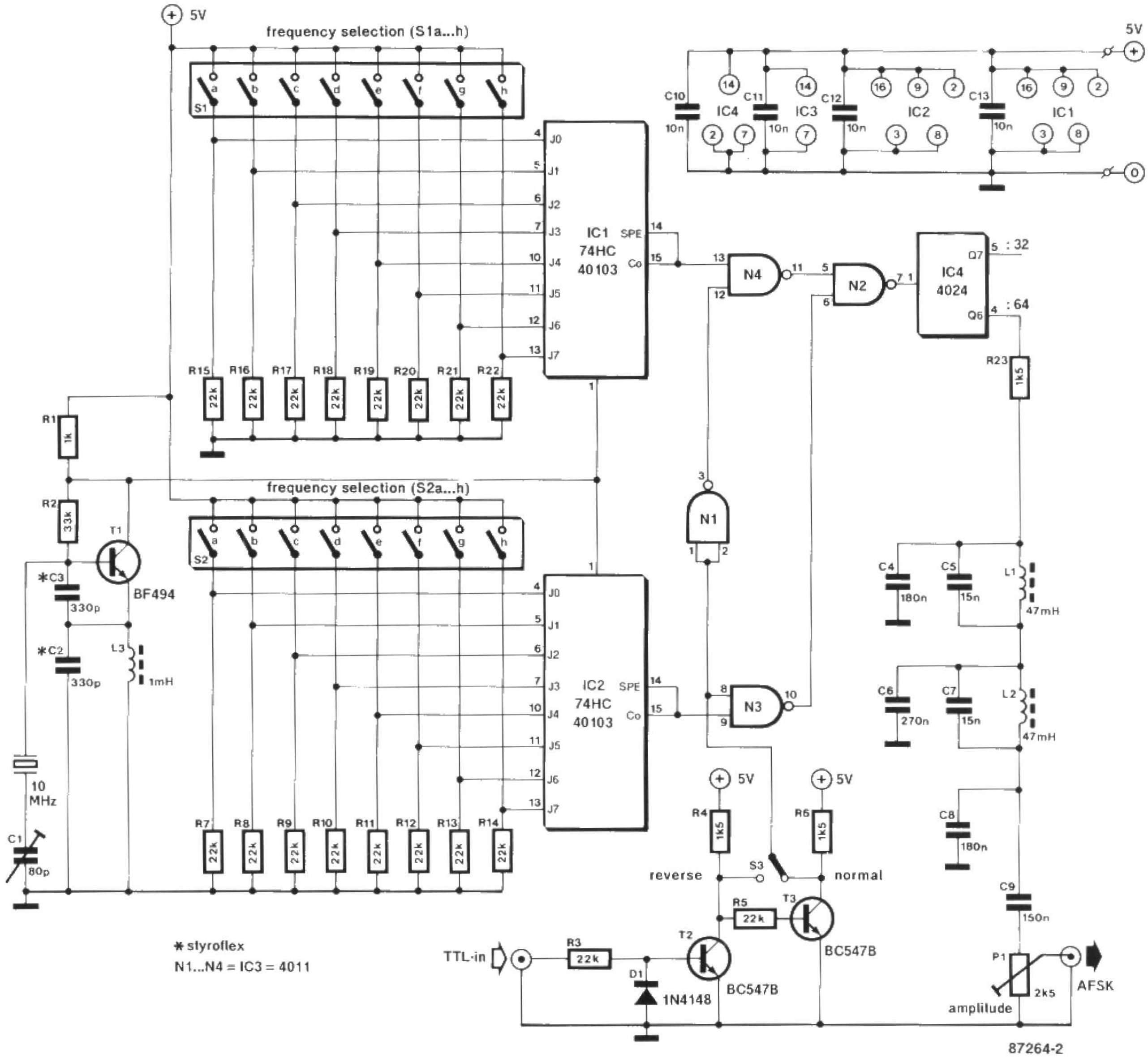

Fig. 15. Circuit diagram of the audio frequency shift generator for RTTY transmitters.

transmitters (VHF), however, it is desired to approach standardized AFSK frequencies as closely as possible. This can be achieved by setting IC_1 and IC_2 to the highest possible divisor, so that the output frequencies can be altered in small steps, and having IC_4 divide by 32 instead of 64 at the cost of a small increase in pulsewidth deviation.

The following example shows how to set the DIP switches in the AFSK generator to obtain mark/space frequencies of 1200 and 2400 Hz. Arithmetically, it makes no difference which of the divisors is defined first. To begin with, there is the fixed divisor at the output (IC_4): 10 MHz divided by 64 equals 156,250 Hz. This, in turn, must be divided down, with divisors between 1 and 256 available (IC_1; IC_2). For instance, in

Parts list
AFSK GENERATOR. CIRCUIT DIAGRAM:
FIG. 15.

Resistors (±5%):
R_1 = 1K0
R_2 = 33K
R_3;R_5;R_7 . . . R_{22} incl. = 22K
R_4;R_6;R_{23} = 1K5
P_1 = 2K5 or 2K2 preset

Capacitors:
C_1 = 80p trimmer
C_2;C_3 = 330p polystyrene (styroflex) 5%
C_4;C_8 = 180n
C_5;C_7 = 15n
C_6 = 270n
C_9 = 150n
C_{10};C_{11};C_{12};C_{13} = 10n

Inductors:
L_1;L_2 = 47mH (e.g. Toko 181LY-473; Cirkit, Bonex)

Semiconductors:
D_1 = 1N4148
IC_1;IC_2 = 74HC40103
IC_3 = 4011
IC_4 = 4024
T_1 = BF494
T_2;T_3 = BC547B

Miscellaneous:
X_1 = 10 MHz quartz crystal (series resonance).
S_1;S_2 = 8-way DIL switch block.
S_3 = miniature SPST switch.
Veroboard as required.

the case of IC_1, the result should be as close as possible to 2400 Hz. Dividing 156,250 by 2400 gives 65.1042, which is rounded off to 65. Due to the internal structure of the 74HCT40103, the set divisor becomes 64 instead of 65. This means that binary code 0100 0000 is applied to inputs $J_0 \ldots J_7$ incl. (J_0=LSB, J_7=MSB). In other words, only J_6 is held logic high.
The decimal values with J_0, J_1, J_2, J_3, J_4, J_5, J_6 and J_7 are 1, 2, 4, 8, 16, 32, 64 and 128 respectively. Any divisor between 1 and 256 can, therefore, be set by closing one or more switches.

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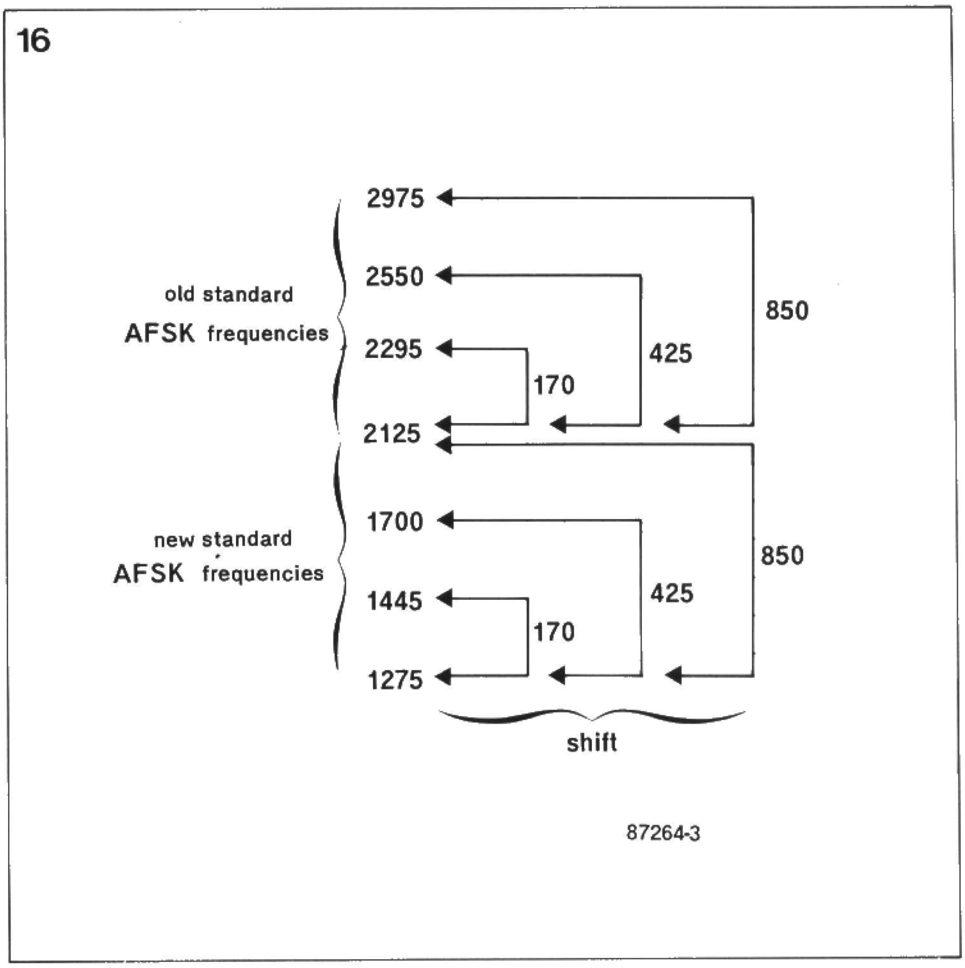


Fig. 16. Overview of 'old' and 'new' AFSK frequencies, and associated shifts.

THE BLACK JAGUAR BJ200 MK2 POCKET SCANNER

An AM/FM 16-channel synthesizer-controlled pocket scanning receiver, the Black Jaguar BJ200 Mk2, is now available from Nevada. The pocket scanner is supplied complete with an instruction manual, a wall-type battery charger, a built-in NiCd battery pack, carrying case, helical rubber acrial, earphone and TNC-to-BNC adaptor to connect an external aerial. The main technical specifications of the scanning receiver are shown in the Table.
One of the remarkable features of the BJ200 is its ability to receive stations on frequencies outside the stated scan ranges. It is, for instance, possible to select, say, 24.5 MHz by entering this frequency on the keyboard. The receiver does not accept it, however, as an upper or lower limit of the scan range.

The outstanding points of the review sample supplied by Nevada are:

1. Given the wideband design of the input stage, sensitivity is excellent on all

Specification

Frequency range	
Band A (HF):	26 – 29.995 MHz in 5 kHz steps
Band B (VHF low):	60 – 88 MHz in 5 kHz steps
Band C (Air & VHF mid):	115 – 178 MHz in 5 kHz steps
Band D (VHF high):	210 – 260 MHz in 10 or 12.5 kHz steps
Band E (UHF):	410 – 520 MHz in 10 or 12.5 kHz steps
Sensitivity (12 dB SINAD for FM; 10 dB S/N for AM; at tune-up)	
FM mode:	0.5 μV for HF and VHF 0.7 μV for UHF
AM mode:	1.0 μV for HF and VHF 1.5 μV for UHF
IF selectivity:	60 dB at ±20 kHz
Spurious and image rejection:	more than 40 dB
Audio output power:	125 mW mininum in 8 Ω
Scan/search speed:	10 channels/s; approx. 45 kHz/s (VHF)
Battery and power consumption:	6 VDC; 5 off NiCd cells. 80 mA squelched, 170 mA at maximum audio
Dimensions (mm):	185(H) × 80(W) × 37(D)
Weight:	approx. 470 g

available bands. For instance, in the 2 m amateur band (144—146 MHz), sensitivity of the review sample was only marginally lower than that of a Type FT227RA synthesizer-driven FM transceiver from Yaesu.

2. Programmed frequencies, channel lock-out, delay and the associated demodulation (AM/FM) are retained in memory even when the battery is completely exhausted. All the programmed settings are available immediately again when the receiver is switched on after charging the batteries.

3. The LC display is well laid-out and clearly legible under virtually all lighting conditions. In the scan mode, only the channel numbers, Lock Out, Delay and AM/FM symbols are displayed to save power. When a station is received, the relevant frequency is shown.

4. Operation is simple and readily learned. The rubber keys have tactile feedback when pressed.

5. The receiver is housed in a rugged ABS enclosure; the keys are recessed and the display is protected by a transparent plastic window. A minor shortcoming is the absence of protective hoods on the two AF output sockets.

6. The built-in battery pack will last for 6 to 8 hours of operation, and can be charged within 12 hours.

7. Frequency ranges provided are useful and definitely the most popular. AM demodulation is available on all bands, and should be of particular interest for VHF airband reception (108—125 MHz). The squelch also works in the AM mode.

The sample BJ200 also exhibited some (mainly minor) deficiencies:

1. The most serious shortcoming of the receiver is its inadequate suppression of harmonics and spurious emissions generated by the built-in synthesizer. When the search mode is used to scan a user-defined frequency range for activity, the receiver will be blocked by its own spurious products which occur at all multiples of 9 MHz. This means that the scanner halts at 27.000 (HF band), 63.000 MHz, 72.000 MHz (VHF low), and so on, right up into the UHF band. The harmonics are so strong as to make reception on and around these frequencies impossible.

2. Spurious products generated by the multiplexed liquid crystal display are particularly troublesome in the HF band (26—30 MHz). Digital noise is picked up on and around 27.0750, 27.215, 27.355, 27.500, 27.640, 27.780 and 27.920 MHz, and a good many other frequencies towards the end of the search range, 29.990 MHz. This interference clearly originates from the LCD because the whirring note produced changes when the Lock Out or Delay function is

activated, and the corresponding LCD symbol lights. Like the previously mentioned synthesizer harmonics, LCD interference halts the scanner indefinitely in the search mode, so that the user is forced to skip the relevant frequency by pressing the UP key.
Fortunately, LCD interference is less severe in the VHF and UHF bands. None the less, it can be heard on some channels as a soft background noise on demodulated speech.

3. When a certain frequency range is scanned for activity (search mode), the receiver halts when the signal strength exceeds the threshold set with the squelch control. In the case of a strong local NBFM signal, scanning is halted before the receiver has reached the actual carrier frequency. This means that it produces unintelligible splatter typically heard when an FM receiver is not properly tuned to a transmission. In this reviewer's opinion, this shortcoming would have been relatively simple to eradicate by controlling the search stop function from the centre tuning output of the FM detector in conjunction with the fieldstrength threshold, thereby ensuring that scanning does not stop until the carrier frequency is reached.

4. Although the receiver was actually observed to switch off its display and audio amplifier on several occasions when the battery was exhausted, the flashing LO BATT symbol did not appear on the LCD as stated in the instruction manual. This is not a serious problem, however, and may have been a fault in a our review model.

5. Considering the use nowadays of many different pre-emphasis standards in NBFM telephony services, a continuously variable tone control would have done better than the two-position selector provided on the BJ200. In some cases, the sound produced lacks brightness, and this is only aggravated by pressing the tone switch, which activates a low-pass AF filter.

6. Local oscillator suppression at the input of the BJ200 is inadequate: a spectrum analyser connected to the aerial socket measured a LO level as high as −35 dBm in the UHF band. Synthesizer sidebands were also clearly observed.

7. Selectivity and image rejection were just about acceptable, although a mobile telephone repeater at a distance of about 2 km was received on 433.920 MHz (in the 70 cm amateur band), while the actual transmit frequency was 443.920 MHz, i.e., 10 MHz higher. The receiver was unable to keep two strong signals from local cellular radio repeaters apart at a channel spacing of 15 kHz.

Conclusions

Notwithstanding the above criticisms, which are mainly inherent to the wideband design, the Black Jaguar BJ200 MK2 is well worth its cost.
The high sensitivity, ruggedness and the well laid-out LCD make the BJ200 a good choice for portable use, although it will function equally well in a fixed location, connected to an external aerial. It is a very compact and reliable receiver with good external finish. The accessories for it are useful and need not be purchased separately, as is often the case with competitive products.

The recommended retail price of the Black Jaguar BJ200 Mk2 scanning receiver is £225. Further information is available from

Nevada • 189 London Road • North End • PORTSMOUTH PO2 9AE. Telephone: (0705) 662145.

NEW COMPUTER SYSTEM ENHANCES TEXTILE PRODUCTION

by Anna Kochan, BEng

An integrated computer system that can warn of production bottlenecks a year ahead and can be used by an operator with no experience of data processing has been on the market since late last year.

Made by McGuffie Brunton Northern[1], the first installation is at the northeast England textile manufacturing mill of the J.H. Walker Company, which collaborated in the design. Walker specializes in sliver pile fabric and jersey fleece.

McGuffie started as a partnership in 1981, became a limited company two years later and developed an annual turnover of £1.8 million by 1986. Its first products were the Trader 25 and Jobber 25 software packages for the wholesale and distribution industries, and make-to-order, finish-to-order and jobbing manufacture, respectively.

These are now installed in more than 250 businesses in the United Kingdom. The new package has evolved from these two earlier versions to suit the requirements of weavers, spinners, dyers, yarn extruders and finishers.

For the past years the firm has been licensed to sell certain ICL computers and it is the ICL System 25 range of business minicomputers that the new package for the textile industry has been chosen to run on.

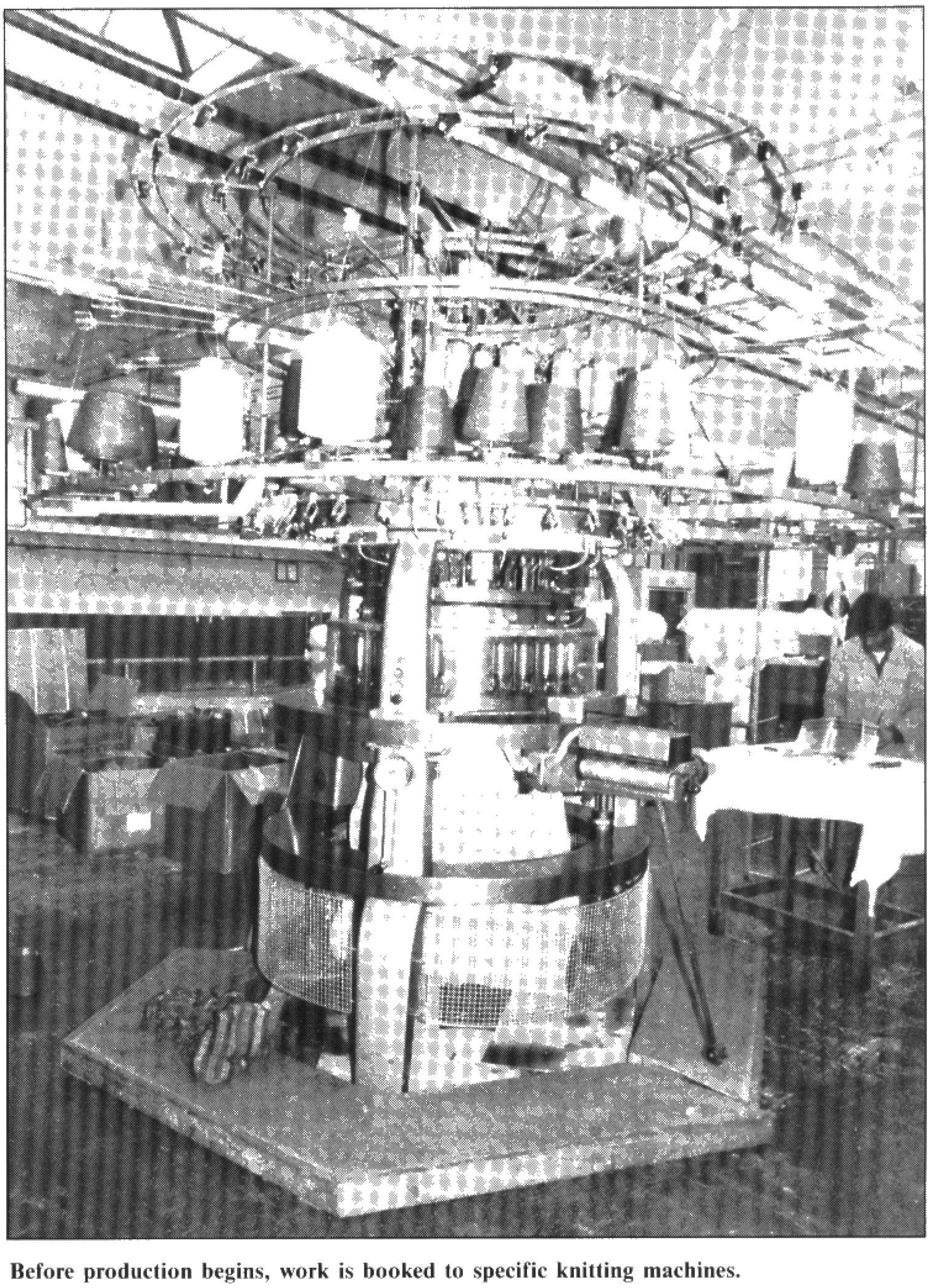
Textile 25, as the new integrated company management control system is known, covers the needs of production control and costing, stock sales orders and financial accounting with comprehensive management reporting.

Total control

The system is modular and can be installed progressively from a modest beginning. The modules are designed in such a way that they can be used by an inexperienced operator and the programs lead the operator through the systems by means of conversational English screen prompting.

Whithin the Textile 25 system, 27 different modules are included and they fall into six broad categories as follows:

* Sales order control: sales order processing, telephone selling option, order intake analysis, production batch-to-sales allocation, despatch control and invoicing.
* Stock control: stock recording, batch control and location control, stock movement audit, stock management and demand forecasting.
* Production control: job and batch estimating, works order documentation, batch costing and labour analysis, production piece tracking, blending and recipe management, bill-off-resource management and requirements planning.
* Buying control: buyer's information database and purchase order process-



Before production begins, work is booked to specific knitting machines.

ing.

* Financial control: sales ledger, sales analysis, purchase ledger, nominal ledger, fixed ledger, fixed assets, payroll and BACS (bankers automatic clearance system) link.
* Data and security link: menu system and data management system.

Avoiding bottlenecks

J.H. Walker uses this new system in the manufacture of a range of high-quality knitted fashion fabrics for the clothing and upholstery trade. The company's two major manufacturing ranges are a jersey and a sliver fabric. The jersey fabric is knitted and stored in a 'grege' (natural) state, and then dyed and finished to individual customer orders. The sliver fabric is knitted and finished in one pass to individual customer requirements. This involves special fibre blending and very precise quality control.

So far, Walker has implemented the first phase of Textile 25 for production, order control and tracking system. A network of factory data collection terminals distributed throughout the production area and connected to the ICL system 25 minicomputer has been installed for this purpose.

As customers' orders are received they are entered on to the computer. The Textile 25 order processing module is specially adapted to handle the many thousands of quality and colour combinations that are possible without the need to store each one. Production batches can be planned and allocated to orders well in advance of start. The requirements planning module highlights potential bottlenecks and material shortages for up to one year ahead, allowing timely action to be taken to rectify the situation.

As work on a production batch is started, it is booked on to a knitting machine. Pieces coming off the machine are then weighed and measured. The system assigns a piece number and prints a bar-coded ticket immediately to identify the piece and accompany it around the factory. As operations are completed, the piece bar code is read and completion of the operation automatically recorded.

Accurate information

At the final inspection, extra information such as quality, net weight and net length are entered via the factory terminal, and a new bar code roll card is printed to accompany the completed piece into store.

Customer despatches are carefully controlled by the system. Each roll card being despatched is bar-code read and checked to be of the correct quality, shade and so on, for the order. Despatch documentation is printed instantly giving the customer full details of roll lengths. Invoicing then follows automatically, quickly and accurately.

At any time the firm can view the current state of a production batch and customer order. Reports highlight over-production and under-production or potential late delivery situations. Comprehensive yield analysis and raw material location control monitors costs and minimises waste.

The system has enabled Walker to improve its customer service levels in terms of meeting delivery deadlines and accurate despatches. Usage of raw materials stocks has been improved and stock holding has been reduced. The system operates 24 hours a day, six days a week and is proving to be reliable and resilient.

In the future, Walker plans to introduce enhancements to its system to cover fibre blending, time and attendance recording and automatic machine motoring.

References:

1. J.H. Walker Ltd, Ravensthorpe Mills, Calder Road, DEWSBURY WF13 3SJ.
2. McGuffie Brunton Northern, The Granary, 50 Barton Road, Worsley, MANCHESTER M28 4PB.

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	880016-4	0.75*	0.11
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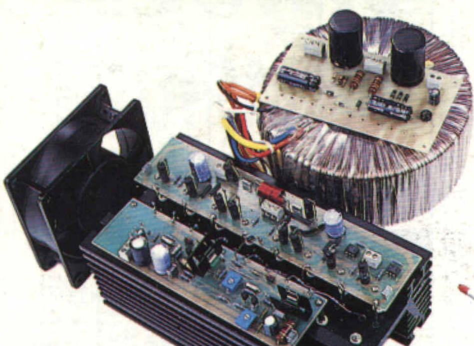

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